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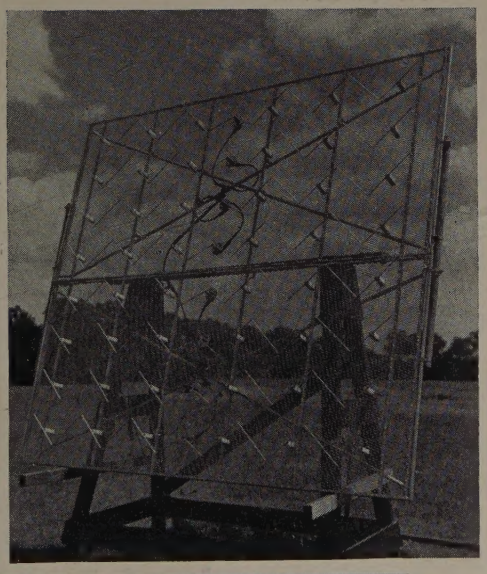
Volume 42 Number 11

TV BOOSTER ANTENNAS



nsmitting antenna
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antenna (right)
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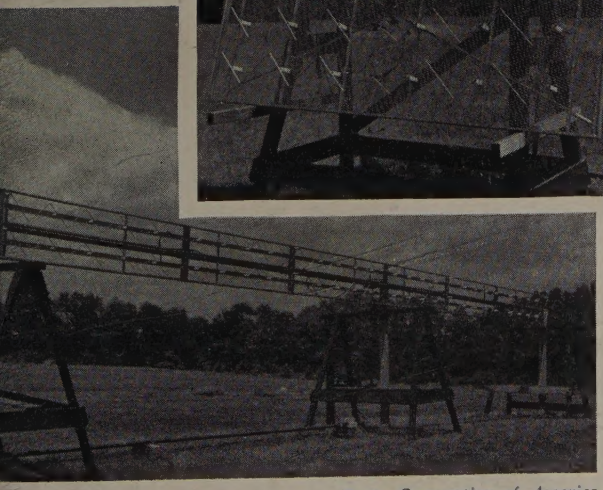

Radio Corporation of America

IN THIS ISSUE

Ferroelectrics and the Dielectric Amplifier
Junction Transistor Design at Higher Frequencies
Rippled Wall and Rippled Stream Amplifier
Stability Considerations in VHF Transistors
Analysis of Linear Systems
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Directional Electromagnetic Couplers
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Abstracts and References

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The Institute of Radio Engineers

THE STANDARD OF COMPARISON

HIGH FIDELITY TRANSFORMERS

FROM STOCK... ITEMS BELOW AND 650 OTHERS IN OUR CATALOGUE B.

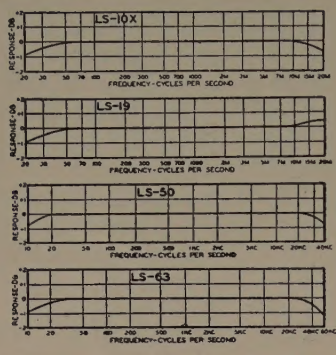
TYPICAL UNITS

LINEAR STANDARD series

Linear Standard units represent the acme from the standpoint of uniform frequency response, low wave form distortion, thorough shielding and dependability. LS units have a guaranteed response within 1db. from 20 to 20,000 cycles.

Hum balanced coil structures and multiple alloy shielding, where required, provide extremely low inductive pickup.

These are the finest high fidelity transformers in the world. 85 stock types from milliwatts to kilowatts.

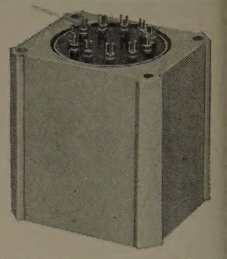


LS-10X Shielded Input
Multiple line (50, 200, 250, 500/600, etc.) to 50,000 ohms . . . multiple shielded.

LS-19 Plate to Two Grids
Primary 15,000 ohms.
Secondary 95,000 ohms C.T.

LS-50 Plate to Line
15,000 ohms to multiple line . . . +15 db. level.

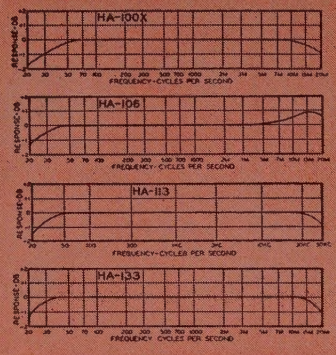
LS-63 P.P. Plates to Voice Coil
Primary 10,000 C.T. and 6,000 C.T. suited to Williamson, MLF, ul.-linear circuits. Secondary 1.2, 2.5, 5, 7.5, 10, 15, 20, 30 ohms. 20 watts.



CASE	LS-1	LS-2	LS-3
Length	3⅛"	4-7/16"	5-13/1
Width	2⅝"	3½"	5"
Height	3¼"	4-3/16"	4-11/1
Unit Wt.	3 lbs.	7.5 lbs.	15 lbs.

HIPERMALLOY series

This series provides virtually all the characteristics of the Linear Standard group in a more compact and lighter structure. The frequency response is within 1 db. from 30 to 20,000 cycles. Hipermalloy nickel iron cores and hum balanced core structures provide minimum distortion and low hum pickup. Input transformers, maximum level +10db. Circular terminal layout and top and bottom mounting.

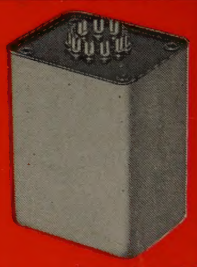


HA-100X Shielded Input
Multiple line to 60,000 ohm grid . . . tri-alloy shielding for low hum pickup.

HA-106 Plate to Two Grids
15,000 ohms to 135,000 ohms in two sections . . +12 db. level.

HA-113 Plate to Line
15,000 ohms to multiple line . . . +12 db. level . . . 0 DC in primary.

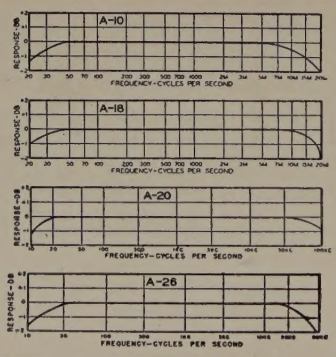
HA-133 Plate (DC) to Line
15,000 ohms to multiple line . . . +15 db. level . . . 8 Ma. DC in primary.



Case	H-1	H-2
Length	2⅛"	3-9/1
Width	1-15/16"	2-13/
Height	3⅛"	3½"
Unit Weight	2 lbs.	5 lbs.

ULTRA COMPACT series

UTC Ultra Compact audio units are small and light in weight, ideally suited to remote amplifier and similar compact equipment. The frequency response is within 2 db. from 30 to 20,000 cycles. Hum balanced coil structure plus high conductivity die cast case provides good inductive shielding. Maximum operating level is +7db. Top and bottom mounting as well as circular terminal layout are used in this series as well as the ones described above.

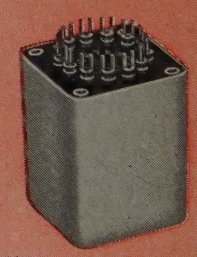


A-10 Line to Grid
Multiple line to 50,000 ohm grid.

A-18 Plate to Two Grids
15,000 ohms to 80,000 ohms, primary and secondary both split.

A-20 Mixing Transformer
Multiple line to multiple line for mixing mikes, lines, etc.

A-26 P.P. Plates to Line
30,000 ohms plate to plate, to multiple line.

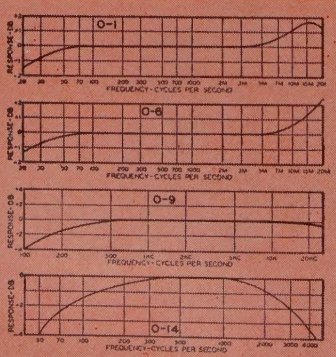


A CASE

Length	1¼
Width	1¼
Height	2"
Unit Weight	½

OUNCER series

UTC Ouncer units are ideal for portable, concealed service, and similar applications. These units are extremely compact . . . fully impregnated and sealed in a drawn housing. Most items provide frequency response within 1 db. from 30 to 20,000 cycles. Maximum operating level 0 db. These units are also available in our stock P series which provide plug-in base. The 0-16 is a new line to grid transformer using two heavy gauge hipermalloy shields for high hum shielding.

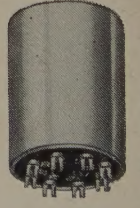


0-1 Line to Grid
Primary 50, 200/250, 500/600 ohms to 50,000 ohm grid.

0-6 Plate to Two Grids
15,000 ohms to 95,000 ohms C.T.

0-9 Plate (DC) to Line
Primary 15,000 ohms, Secondary 50, 200/250, 500/600.

0-14 50: 1 Line to Grid
Primary 200 ohms, Secondary .5 megohm for mike or line to grid.



OUNCER CASE

Diameter	⅞"
Height	1-3/
Unit Weight	1 oz.

SPECIAL UNITS TO YOUR NEEDS

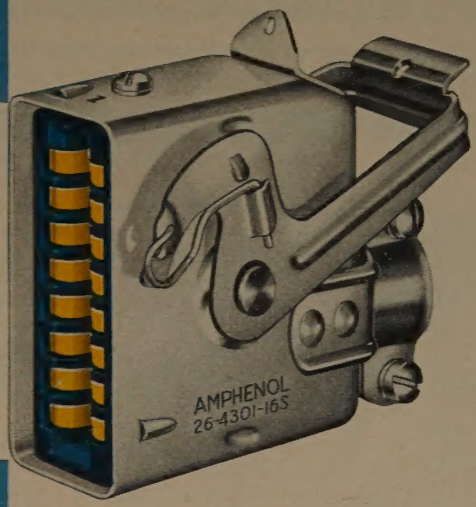
If you manufacture high fidelity gear, send your specifications for prices.

UNITED TRANSFORMER C

150 Varick Street, New York 13, N. Y. EXPORT DIVISION: 13 E. 40th St., New York 16, N. Y. CABLES: "AR

AMPHENOL

the famous BLUE RIBBON CONNECTORS



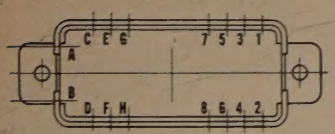
now available with Panel & Latch-Lock shells!

AMPHENOL's crack engineering team have completely redesigned the famous *Blue* RIBBON connectors and the results will be of interest every company engaged in electronics.

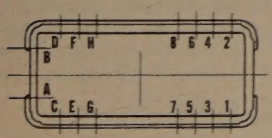
The basic improvement has been the design of a new method of polarization. Instead of conventional guide pin and bushing polarizatio male and female *Blue* RIBBON connectors are now mated by means of proper matching of the barrier heights betwen the contacts. Following this first important improvement AMPHENOL's engineers designed a complete line of front panel shells and cable-clamp latch-lock cans to the connectors—a step forward that makes the versatile *Blue* RIBBONs even more useful to the electronics industry.

What are the advantages of these design changes to you? Barrier polarization allows increased contact spacing without extendi the overall length of the connector—mismating is impossible. Front panel shells and latch-lock cans are available for *Blue* RIBBONs in a wide variety of keying arrangements, making possible the mounting of large numbers of connectors side by side without the possibility of connector mismating by untrained personnel. Latch-lock types may be safety-wired and the cans are available with either end or side cable outlets.

Alternate Keying Positions



Receptacle Front Shell



Plug Front Shell

lternate keying on BLUE *RIBBON shells rovide positive insurance against mismat- g in side-by-side connector mountings.*

Complete details on all AMPHENOL Blue RIBBON connectors in CATALOG R1

AMPHENOL

AMERICAN PHENOLIC CORPORATION *chicago 50, illinois*

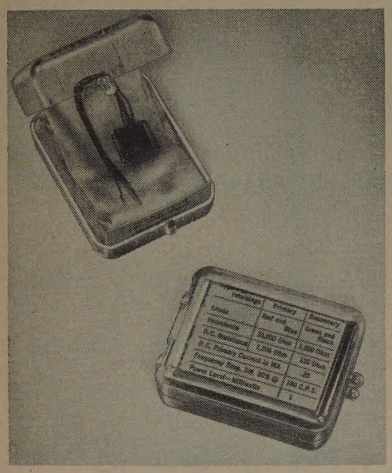
In Canada: AMPHENOL CANADA LIMITED, Toronto

NEWS and NEW PRODUCTS

November 1954

Transistor Transformer

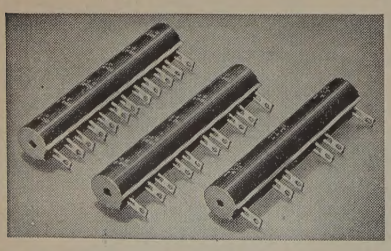
The design and production of a transistor transformer with a wide variety of industrial applications has been announced by **Telex, Inc., Acoustic Div.,** Telex Park, St. Paul 1, Minn. The component measures $\frac{1}{2} \times \frac{7}{16} \times \frac{3}{8}$ inches in size, and the transformer comes in input, interstage and output models, #8641 Transistor Input, #8642 Transistor Inter-Stage and #8643 Transistor Output.



Impedance of the Transistor Input's primary coil is 200,000 ohms, with current in the primary coil of 0 milliamperes (dc) and impedance in the secondary coil of 1,000 ohms; frequency response is down 3 db at 200 cps per second from 1,000 cps n the input stage, and the maximum power at 200 cps is 0.5 milliwatt. Figures on the Transistor Inter-Stage, in the same order as above, are: 20,000 ohms, 0.25 ma, 1.00 ohms, down 3 db at 100 cps and maximum power at 100 cps is 1 milliwatt. For the Transistor Output the figures are: 1,000 ohms, 2.0 ma, 100 ohms, down 3 db at 120 cps and maximum power at 120 cps is 15 milliwatts.

Encapsulated Precision Resistors

A new "grouped" model of the recently released Riteohm® encapsulated precision resistors has just been announced by **Ohmite Manufacturing Co.,** 3664 Howard St., Skokie, Ill.



The new grouped type units are made up of from two to six individual, lug-type Riteohm resistors, arranged end-to-end in a one-piece, molded resin body. (Illustration shows three, five and six resistor units.)

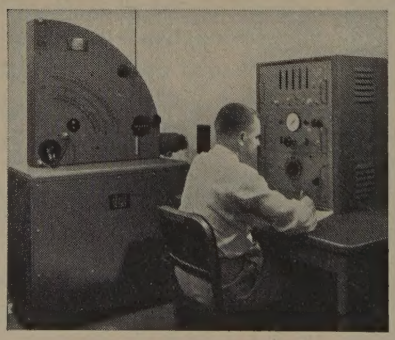
A special Ohmite-compounded resin completely encloses the individual resistors and gives maximum protection against mechanical damage and ambient conditions. The coefficient of expansion of the resin closely matches that of the other materials of the unit.

New Grouped Type Riteohms are available in lengths up to approximately three inches. The individual resistor units in the group can be obtained in wattage sizes of $\frac{1}{4}$, $\frac{1}{2}$, and 1 watt (125°C ambient), and in resistance values up to a maximum of 1,950,000 ohms. Tolerances as close as ±0.1 per cent are available. Request Bulletin No. 145.

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

High-Speed Framing Camera

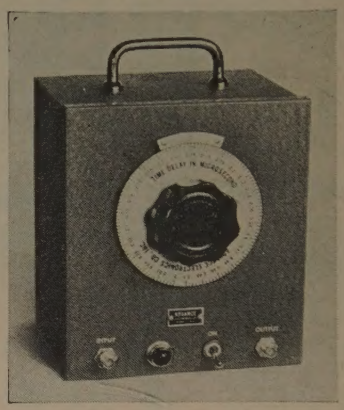
Producing high-resolution photographs as rates as high as 2.4 million per second, the new Model 222 Framing Camera developed by **Beckman & Whitley, Inc.,** 963 San Carlos Ave., San Carlos, Calif., serves the research needs of workers in combustion, corona-discharge, explosion, plastic- and elastic-deformation, and shock-wave phenomena. A similar but lower-speed Model 189 covers the range up to 1.2 million frames per second.



An image of the subject matter is focused on the stainless- or alloy-steel turbine-mirror through a highly-corrected 24-inch achromat. This image is related 1-to-1 to the film plane on a 30-inch radius through 25 pairs of adjustable achromats. Plano-concave, field-flattener lenses are located at each frame position within 0.002 inches of the film emulsion surface. A total of 25 frames per run can thus be exposed on stationary perforated 35-mm film.

Time Delay Network

Type 303 continuously variable Time Delay Passive Network developed by **Advance Electronics Co., Inc.,** 451 Highland Ave., Passaic, N. J., is suitable for use as a time delay matching device in television systems, variable time delay for pulses, or precision measurement of small time intervals. The rise time is less than 7 per cent of the time delay at any point and the amount of overshoot is less than 2 per cent.

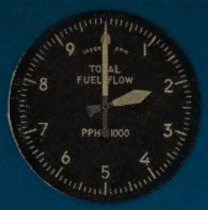


This instrument consists of an input amplifier, an output amplifier and a continuously variable delay line which may have ten variations. The shortest one is 0 to 0.05 microsecond and the longest one is 0 to 0.8 microsecond. The input impedance is 1 megohm shunted with 20 $\mu\mu$f. The output impedance is 1000 ohms shunted with 15 $\mu\mu$f. Both the input and output impedance can be made lower by shunting the terminals with proper resistors. Both amplifier stages have 15 mc bandwidth and 10 volts peak-to-peak maximum signal level. Accuracy can be maintained within ±2 per cent after calibration Type 303 costs $85.00 each plus the delay line.

"C" Bank Wavemeter

Amerac, Inc., Wenhan, Mass., announces production of the Model 230 "C" Band Wavemeter. This is a coaxial line type instrument, covering the frequency range from 3,500 mc, to 6,500 mc, by either the transmission or absorption method.

(Continued on page 18A)



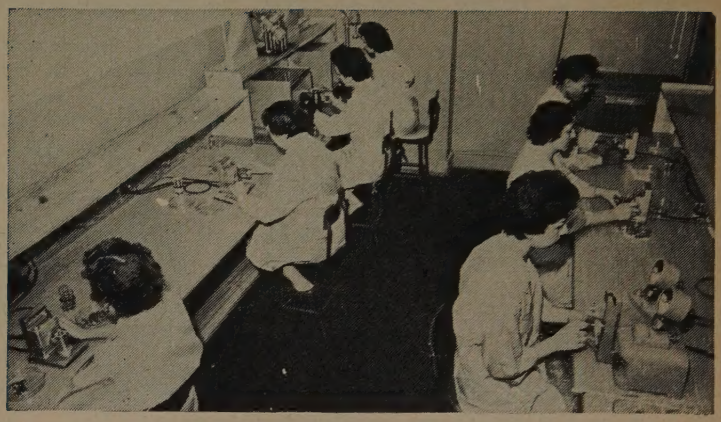
FUEL TOTALIZING INDICATOR

FUEL FLOW TRANSMITTER

DUAL FUEL FLOW INDICATOR

Electronic control devices are among the many instruments Ketay manufactures for aircraft, missiles, marine, ordnance, and civilian application. Special designs to meet the limitations of space and operating conditions.

tay



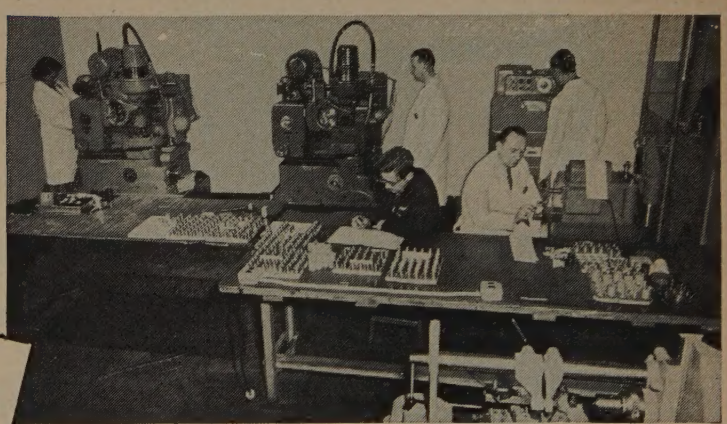
Inspection of all parts, sub-assemblies, as well as completed instruments, is a fetish at Ketay. For instance, all bearings are inspected with specially developed equipment in air conditioned work space. This is just one of the many techniques that assure maximum performance when Ketay units are specified.

electro-mechanical devices some of which are illustrated on these pages. Currently, original Ketay developments are providing instrument performance far above present standards—many of which were set by earlier Ketay developments.

Ketay successfully applies its production facilities and experienced research personnel to specific problems for the leaders in automatic control.

Your interest will be well served by learning fully of the products and services you may obtain from Ketay.

Write, today, for descriptive specifications of those types of Ketay instruments which may be applied to your designs.



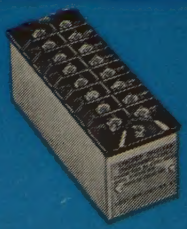
Precision of manufacture is vital in every Ketay unit. To assure continuing reliability for its products, Ketay employs the very latest facilities and techniques. Typical is this "gear room" where modern gear cutting machinery produces gears to the finest of tolerances.

Pacific Division:
12833 Simms Avenue, Hawthorne, Calif.

Kinetix Instrument Div., Pacific Div.
Research & Development Div.

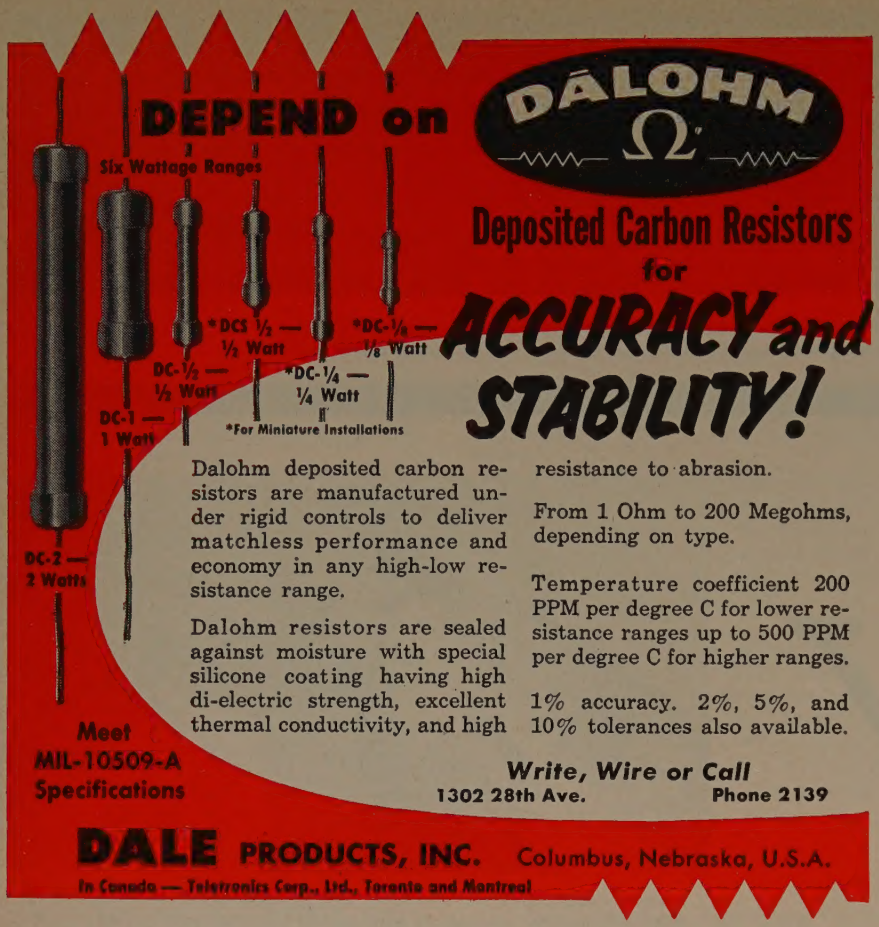
TACHOMETER GENERATOR

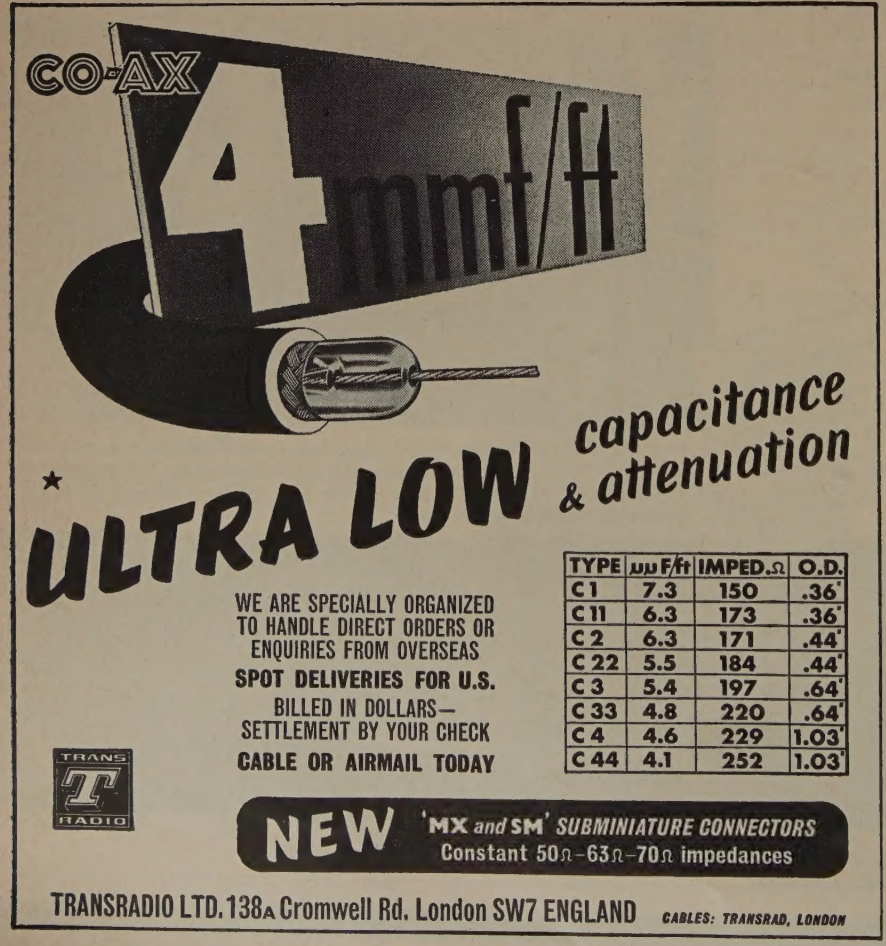
PANCAKE SYNCHRO GYRO PICKOFF



SYNCHRO OVERLOAD TRANSFORMER

Many specialized units have been designed by Ketay engineers and are in quantity production. Custom engineered units for specific application are also available.


DEPEND on
DALOHM Ω
Six Wattage Ranges
Deposited Carbon Resistors
for
ACCURACY and
STABILITY!
*DCS ½ — ½ Watt
*DC-⅛ — ⅛ Watt
DC-½ — ½ Watt
*DC-¼ — ¼ Watt
DC-1 — 1 Watt
*For Miniature Installations
DC-2 — 2 Watts
Dalohm deposited carbon resistors are manufactured under rigid controls to deliver matchless performance and economy in any high-low resistance range.
Dalohm resistors are sealed against moisture with special silicone coating having high di-electric strength, excellent thermal conductivity, and high resistance to abrasion.
From 1 Ohm to 200 Megohms, depending on type.
Temperature coefficient 200 PPM per degree C for lower resistance ranges up to 500 PPM per degree C for higher ranges.
1% accuracy. 2%, 5%, and 10% tolerances also available.
Meet MIL-10509-A Specifications
Write, Wire or Call
1302 28th Ave. Phone 2139
DALE PRODUCTS, INC. Columbus, Nebraska, U.S.A.
In Canada — Telerronics Corp., Ltd., Toronto and Montreal


CO-AX
4 mmf/ft
★ ULTRA LOW capacitance & attenuation
WE ARE SPECIALLY ORGANIZED TO HANDLE DIRECT ORDERS OR ENQUIRIES FROM OVERSEAS
SPOT DELIVERIES FOR U.S.
BILLED IN DOLLARS— SETTLEMENT BY YOUR CHECK
CABLE OR AIRMAIL TODAY
TYPE | μμF/ft | IMPED.Ω | O.D.
C 1 | 7.3 | 150 | .36"
C 11 | 6.3 | 173 | .36"
C 2 | 6.3 | 171 | .44"
C 22 | 5.5 | 184 | .44"
C 3 | 5.4 | 197 | .64"
C 33 | 4.8 | 220 | .64"
C 4 | 4.6 | 229 | 1.03"
C 44 | 4.1 | 252 | 1.03"
TRANS T RADIO
NEW 'MX and SM' SUBMINIATURE CONNECTORS
Constant 50Ω–63Ω–70Ω impedances
TRANSRADIO LTD. 138A Cromwell Rd. London SW7 ENGLAND
CABLES: TRANSRAD, LONDON

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

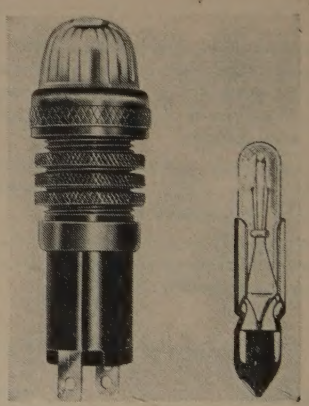
(Continued from page 14A)

It features: (1) a precision-ground lead screw, which helps give accuracy of measurement, (2) a cavity body made from a solid block, giving good mechanical stability, (3) the use of Invar in the line displacement portion, affording a high frequency stability throughout the temperature range of 10°C to 40°C, (4) Tri-plating of all RF surfaces.

Type "N" constant impedance coaxial connectors are used for both transmission and absorption inputs; BNC or UHF coaxial fitting for external video connection; Power-handling capacity by absorption method is from 0.5 milliwatt to 1 watt maximum; Power-handling by transmission method is from 1 milliwatt to 25 watts (peak power); Approximate loaded Q is 2,500; Cabinet size is 8 wide×6½ deep×5 inches high. Net weight is 4¾ pounds.

Indicators For Slide Base Lamps

The Dialco line of pilot light assemblies produced by **Dialight Corp.**, 60 Stewart Ave., Brooklyn 37, N. Y., now includes a new series designed to employ any of the telephone type incandescent bulbs which have slide bases. No extractors or other tools are required.



This new series offers a wide choice of screw and friction fitted caps with flat and convex lenses of ⅜ and ½ inch diameter. For wide angle visibility, a unique screw cap with stovepipe lens (molded of glass or plastic) is fitted over a lamp brought unusually far forward in the panel.

Mounting is done in a single 9/16 inch hole. Adjacent units may be placed as close as ¾ inch (center to center).

These pilot lights may be had with soldering terminals or binding screws. Samples for design purposes will be submitted at once and without charge if requested on company letterheads. In addition, the following literature on Dialco pilot light assemblies is available: Catalogs L-151, L-131, and L-154.

(Continued on page 32A)

FIRST AND FOREMOST IN MINIATURIZATION

IN THE "SQUEEZE-PLAY" OF HEAT & PRESSURE— HERMETIC'S A-N CONNECTORS WON'T LEAK!

Hermetic's Vac-Tite* Glass-to-Metal seal guarantees performance in HS 2, HS 6 connectors.

Tough instrumentation problems often dictate use of HS 2 (AN 3102) receptacles and HS 6 (AN 3106) plugs. Superior performance is wanted and superior performance you'll get with Hermetic:

- Vacuum tight (Mass spectrometer proven)
- Arc-resistance of glass
- High-temperature operation
- Corrosion resistance
- 100% moisture and pressure resistant
- Shock and vibration proof
- Equivalent to MIL-C-5015

In addition to our standard line of HS 2 and HS 6 connectors, special units with particular plating requirements, varied flange style, and extra-high pressure resistance, etc., are being designed and manufactured to meet the most specialized needs.

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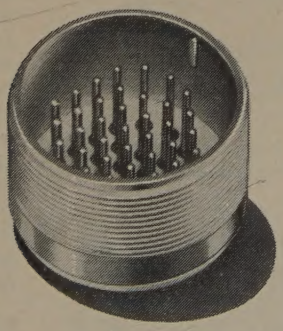
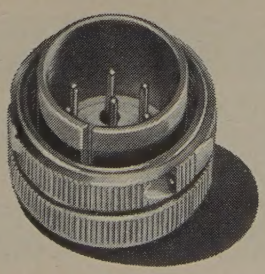
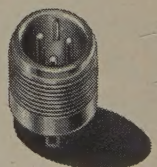
***Vac-Tite is HERMETIC'S new, vacuum-proof, compression-construction, glass-to-metal seal.**







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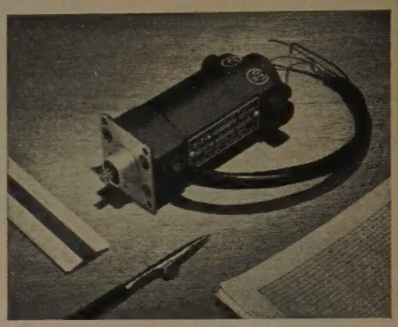
News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 18A)

Actuator Motor

A compact motor for intermittent duty, the new Dalmotor SR-43 manufactured by **Dalmotor Co.,** 1360 Clay St., Santa Clara, Calif., suggests itself for such applications as low- or high-speed, linear- or rotary-actuator mechanisms. The new unit incorporates an automatic clutch brake which applies 2 ounce/inches of brake torque to the load, and disengages the armature automatically when power is removed.



Totally enclosed, the unit turns up 20,000 rpm at 1¾ ounce/inches torque with 3.8 amperes of 28-volt dc power. The locked-rotor torque is 10½ ounce/inches. Rotation is reversible and various shaft arrangements including plain shafts, splines, keyways, and gears to specification can be provided. Electrical connections are by integral shielded leads which can be of various lengths as specified.

Convertible Plastic-Welding Generators

The full normal capacity range of plastic-welding operations for packaging and specialty manufacturing is covered by newly-designed high-frequency Ectrotherm generator line available from **Electronic Processes Corp.,** 1197 San Antonio Rd., Los Altos, Calif. Basic unit, (shown) hasall operating sub-assemblies mounted on sliding-drawer type chassis units for immediate change-over of power-supply and oscillator units. Besides simplifying service and maintenance, this permits on-the-job model conversion. In its higher-powered type, designated Model 253, the generator has an input power of 4,000 dc watts, which provides for average weld lengths of 20 inches although variations in specific conditions permit variations in this capacity up to a maximum of 60 inches on 0.020 vinyl. Operating at an output frequency of 27 mc the Model 253 weighs 113 pounds.

Fitted out with the lower-powered subassemblies, the unit is designated as the Model 153, with an input power of 2,000 dc watts, alternative output frequencies of 27 or 41 mc and a net weight of 94 pounds. This type is rated for average weld lengths of 20 inches, although, depending upon

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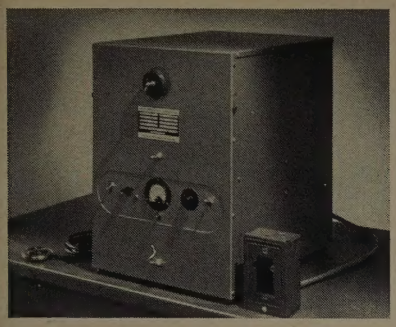
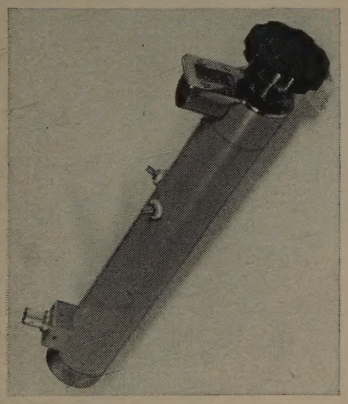


particular conditions, maximum weld lengths of 36 inches on 0.020 vinyl have been performed.

Both models operate from 60-cps, single-phase power, Model 153 using 115-volt supply; Model 253 using 230-volt. They have identical dimensions of 15 by 24 by 21 inches.

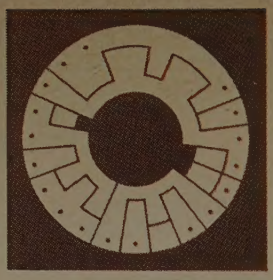
Rocket Tube Cavity Oscillator

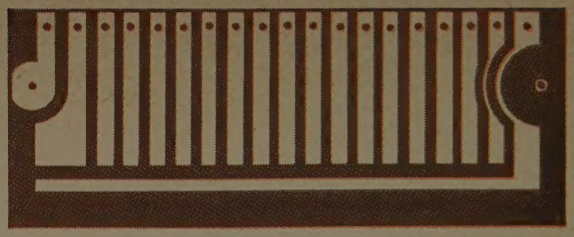
Amerac, Inc., 116 Topsfield Rd., Wenham, Mass., announces the design and production of their #192A Rocket Tube Cavity Oscillator, for microwave signal generation. This is a small-size, co-axial line cavity oscillator, employing the Sylvania uhf planar triode tube.



The unit features a single knob control and utilizes fixed feedback. Designed with emphasis on minification, the #192A is 8 inches long $\times 1\frac{5}{16}$ inches in diameter, and weighs 1 pound, 3 ounces, (including tuning mechanism) or may be supplied as a cavity alone, 4 long $\times 1\frac{5}{16}$ inches weighing $\frac{3}{4}$ pound. It is available at frequencies from 1,000 to 4,000 mc with a tuning range of 400 mc. Pulse operation is 2,000 peak volts with a peak pulse power of 200 watts. CW operation is 175 volts with an average output power of 100 mw. Tuning accuracy is ± 0.1 per cent with regulated power supply. Output connection is BNC or type N jack. It has a Root counter for calibration. The cavity comes complete with tube and can be furnished with a regulated supply and frequency calibration curves.

(Continued on page 130A)





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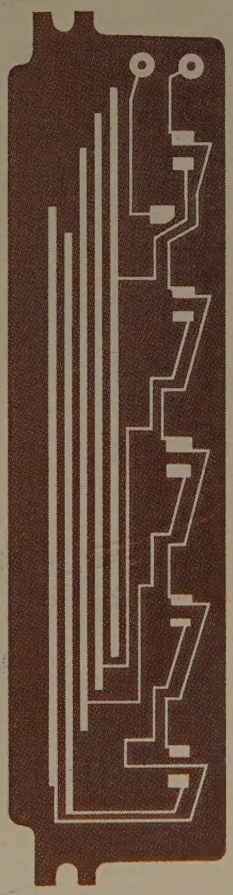

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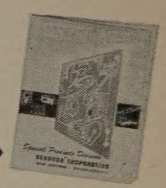

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Suggestions on Content and Presentation

IN ACCORDANCE with policy determined by the Editorial Board, PROCEEDINGS papers must describe sound and important work in a manner which is intelligible to engineers working in other fields. This is so that the PROCEEDINGS can serve not only to provide reference material for specialists but also to keep radio engineers generally informed on progress in the various branches of radio engineering and electronics. It is believed that these objectives are not conflicting and that clarity of presentation is welcomed as much by the expert as by the casual reader. A clearly written paper, moreover, will secure for the author the largest possible audience. The following suggestions as to form and organization are to assist the author in presenting his material to the best advantage.

The paper will ordinarily consist of a Title, Summary or Abstract, Introduction, Main Section and Discussion, in that order. The summary at the beginning, rather than a list of conclusions at the end, has been adopted as the IRE standard form. The roles of these parts of the paper in increasing the clarity of presentation will be discussed in the following paragraphs.

The Title

The title should be short if possible but not so short that the subject will not be indicated clearly. The author should remember that a paper is indexed by significant words in the title and that many readers of the PROCEEDINGS select the papers they read from the table of contents. The title, therefore, should be carefully chosen. Six to eight words is the usual maximum length.

The Summary or Abstract

This section should be at the beginning of the paper because here it leads naturally into the introduction. It is helpful for the reader to be informed briefly of the subject, method, and results before the background of the work is reviewed in the introduction to follow. The summary should be really informative and should outline the essentials of the paper within the compass of about 200 words or less. It should be meaty enough to serve as a complete abstract for such services as Science Abstracts or the Abstracts and References section of the PROCEEDINGS. The author should ask himself if he is willing to accept his abstract, without the rest of the paper, as a fair statement of what he has accomplished. The problem should be clearly stated, the method of attack outlined, and numerical results, where possible, included. In general neither references nor equations should appear in the summary. The essence of a good abstract is a concise statement in plain English of just what has been done and why.

An example of a good abstract is the following one of the paper, "Some New Developments in High Fidelity Loudspeakers," by H. F. Olson and John Preston, which appeared in the 1954 Convention Record, part 6, page 8.

> Three direct-radiator-dynamic-loudspeaker mechanisms have been developed for high-fidelity sound reproduction. These units are of three sizes, namely, 8-, 12-, and 15-inch mechanisms. The improvements include smoother response frequency characteristics, broader directivity patterns, improved transient response, and lower nonlinear distortion. These characteristics have been obtained by employing the proper configuration for the cone, a special pulp for the material of the cone and a damping ring in the outside suspension of the cone. In the case of the 15-inch duo cone loudspeaker an additional feature, namely, a series of conical domes cemented to the low-frequency cone. The high-frequency response is improved by these domes because of the smaller solid angle into which the high-frequency cone operates. The directivity pattern is broadened due to the diffraction effects produced by the domes.

Contrast the above with the following, which is typical of many so-called summaries.

"The design of high-fidelity loudspeakers is discussed and a new line of loudspeakers is described." The reader of such an abstract, if it appears by itself, has no idea whether it will be worth his while to look up the original paper or not. If it appears at the beginning of the paper, it gives the reader no information as to the aspects of the problem considered, the method of attack or the results. The general reader is unlikely to dig into the body of a paper without clearer indication that he will find something of interest.

The Introduction

The purpose of the introduction is to orient the reader with respect to the problem. He should be told how the new work ties in with the old and in what respect the results are new. There is no need for the introduction to be a long scholarly study. The important relationships

can be pointed out in a few words if the author really understands the background of his work. And it helps greatly to make a paper readable if the author will state the purpose of the work again in plain English, before he plunges into the technical details.

The writer of a paper should be scrupulously careful in his introduction to give references to all work on which his paper depends directly or significantly.

The Body of the Report

Here, the writer should avoid following a stereotyped form and should bear in mind constantly that his object is to communicate information effectively to the reader. Even other workers in the same field appreciate clear indications of the line of thought that is being followed, and frequent guideposts are essential to non-specialists who want to understand the general nature of the work and its significance but do not want to go into the mathematical details.

In work that is essentially mathematical, it is most helpful to the reader to carry along with the mathematics a physical picture of the successive stages through which the work is being carried. It is frequently advantageous to put long, purely mathematical derivations in appendices to avoid interrupting the main train of thought.

The Discussion

After the work has been described and the results have been given, the significance of the work should be carefully considered, usually in a separate section. In this part of the report many of the conclusions given in the summary will be developed. The Discussion is of greatest importance, since any work that is worth reporting deserves careful appraisal. Failure to state the conclusions clearly is perhaps the commonest fault in writing a scientific paper.

Figures and Tables

It is very helpful to readers of all classes if tables, figures, curves, etc. can be arranged so that they are self-explanatory and can be used with a minimum of reference to the text. The quantities and units used in plotted curves should be given clearly, and the captions worded to convey as much information as possible to the engineer who is merely leafing through his copy of the PROCEEDINGS. If his attention is caught by curves that mean something to him, he is much more likely to read the paper. A caption such as, "Plot of x_0/y versus s for various values of the parameter n," will not enlist the interest of many readers. It is better to say, "Axial sound-pressure curves of electrostatic loudspeaker for various diaphragm thicknesses."

INSTRUCTIONS ON SUBMITTING PAPERS

The review and publication of papers will be seriously delayed if manuscripts and illustrations are not submitted in complete form. By complying with the following rules, the author can avoid delays, reduce the likelihood of errors, and improve the appearance of his paper.

Material Required

The following material should be submitted to the Managing Editor, Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y.

1. Three copies of the paper for review purposes, each copy to include illustrations, a separate list of captions, and an abstract not exceeding 200 words.
2. The master set of original illustrations (not copies) for printing must be submitted with the paper.
3. A photograph and biography (125 words) of each author must be submitted with the paper.

Preparation of Manuscripts

Manuscripts should be typed on one side of the sheet only and double-spaced. All illustrations should be referred to in the text by figure number. All footnotes should be numbered.

Footnote references must be complete. Authors' initials, title of article, volume and page numbers, and month of publication should be included. The complete reference should be as follows:

For a periodical: R. N. Hall, "Power rectifiers and transistors," PROC. I.R.E., vol. 40, pp. 1512–1518; November, 1952.

For a book: W. A. Edson, "Vacuum Tube Oscillators," John Wiley and Sons, Inc., New York, N. Y., pp. 170–171; 1948.

Abbreviations should be made in accordance with "Standards on Abbreviations of Radio-Electronic Terms, 1951," PROC. I.R.E., vol. 39, no. 4, p. 397; April, 1951.

Preparation of Illustrations

Drawings for printing must be in black ink on white paper or on tracing cloth. Photographs must be glossy prints. The size should not exceed $8\frac{1}{2}$ by 11 inches.

It should be borne in mind that most illustrations will be reduced in size to a $3\frac{1}{2}$-inch column width when printed. It is particularly important, therefore, that all letters, numbers, and lines be drawn large enough and heavy enough to remain legible after reduction.

Drawings with typewriting on them are not acceptable. All information to be reproduced must be hand lettered in ink.

Graphs should be drawn with only the major coordinate lines showing. A chart containing a large number of closely spaced lines will not reproduce legibly.

Captions, if they are included on the drawing itself, should not appear within the area to be reproduced. They should be placed under the illustration.

Graphical symbols should be drawn in accordance with "IRE Standards on Graphical Symbols for Electrical Diagrams, 1954," PROC. I.R.E., vol. 42, no. 6, p. 965; June, 1954.

—*The Editorial Board*

Ferroelectrics and the Dielectric Amplifier*

W. P. MASON†, FELLOW, IRE, AND R. F. WICK†

Summary—**Ferroelectric materials are the dielectric analog of ferromagnetic materials. Their uses parallel those of ferromagnetic materials in such applications as high permeability materials, magnetostrictive transducers, magnetic amplifiers, and magnetic information storage devices.**

A review is given of the measured properties of single crystals to provide a background for the application of ceramics in capacitors and transducers, and their application in dielectric amplifiers. It is shown that the properties of ceramics can be stabilized by introducing amounts of lead and calcium titanate. A long-time aging effect can be reduced by a proper heat treatment. With these stabilized ceramics, applications such as electromechanical filters and delay lines can be satisfied.

The theory of the dielectric amplifier is discussed and it is shown that stable amplifiers can be obtained by using the properties of single crystals. Amplifiers having carrier frequencies of 200 kc, signal frequencies of 0 to 7,000 cycles, and gains of 12 db, have been realized. These operate with single crystals below the Curie temperature and are stable over a wide temperatue range.

Introduction

DURING the last 10 years, ferroelectric materials have been widely used in such devices as high dielectric constant capacitors and electromechanical transducers and their application has been considered for such devices as dielectric amplifiers and information storage devices. Their uses parallel those of ferromagnetic materials in such applications as high permeability materials, magnetostrictive transducers, magnetic amplifiers, and magnetic information storage devices. In fact, the name ferroelectric signifies that these materials reproduce in dielectric form the properties of ferromagnetic materials. Another name, Seignette electric materials, has been applied to this class of materials by the Europeans in honor of the first ferroelectric to be discovered, Seignette or Rochelle salt. However, the name ferroelectric has been very widely standardized in this country.

Although about 25 members of this group have been discovered, the only two members that have been widely used in the ferroelectric region are Rochelle salt and barium titanate. Of these the latter has the advantage that its properties are less sensitive to temperature. Furthermore, a ceramic can be prepared by sintering together large numbers of small single crystals, which has many of the characteristics of the single crystal. For the condenser and transducer applications, ceramic barium titanate, with the additions of certain other titanates, are used in practical embodiments. For dielectric amplifiers and storage devices, single crystals have many advantages. To understand the properties of the ceramics it is necessary to know the properties of single crystals and, hence, a review of the measured properties of the single crystal is given in the next section.

* Decimal classification: R281×R363. Original manuscript received by the IRE, July 27, 1954.

† Bell Telephone Labs., Inc., Murray Hill, N. J.

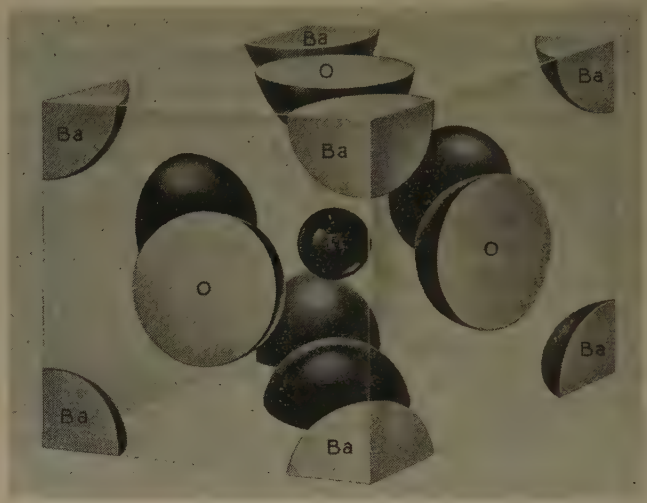

Fig. 1—Block drawing of structure of unit cell of barium titanate crystal.

Properties of Single Crystals

Barium titanate $BaTiO_3$ crystallizes in the cubic system above 120 degrees C. and has the perovskite structure. Fig. 1 shows a drawing of the arrangement of atoms in the unit cell. Barium atoms occupy the eight corners, oxygen atoms occupy the face centered positions of the six faces, and the titanium atom is at the center of the unit cell. Since each barium atom is shared among 8 unit cells, and each oxygen between two cells, the chemical formula is $BaTiO_3$. Fig. 2 shows a drawing

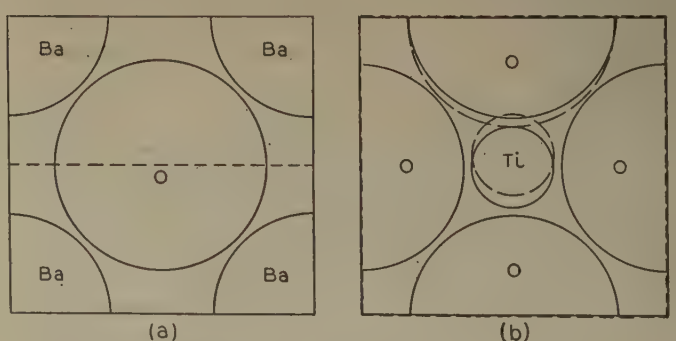

Fig. 2—End view and cross section of $BaTiO_3$ unit cell.

which is more nearly to scale since the atoms come very close together. The top figure (a) shows a unit face. If we cut the cell down the dotted line, the lower figure (b) results. Above 120 degrees C. the atoms occupy the solid line positions, but if the temperature is lowered be-

low the Curie temperature of 120 degrees C., there is considerable X-ray evidence[1] that the titanium moves up toward one of the oxygens and this oxygen moves down to meet it. Slight displacements may also occur in the other four oxygen positions. Since the titanium has a positive charge and the oxygen a negative charge, the centers of gravity of the positive and negative charges no longer coincide and the cell has a spontaneous dipole moment. This rearrangement of atoms causes a distortion in the unit cell and it expands by $\frac{2}{3}$ of 1 per cent in the direction of polarization, which is called the ferroelectric axis, and contracts by $\frac{1}{3}$ of 1 per cent for directions perpendicular to this. This change of dimension is shown in more detail in Fig. 3,[2] and it is seen that a discontinuous separation in dimensions occurs at 120 degrees C. and this separation continues to increase as the temperature decreases toward 20 degrees C. This change in dimensions is the source of the electrostrictive effect discussed in the next section.

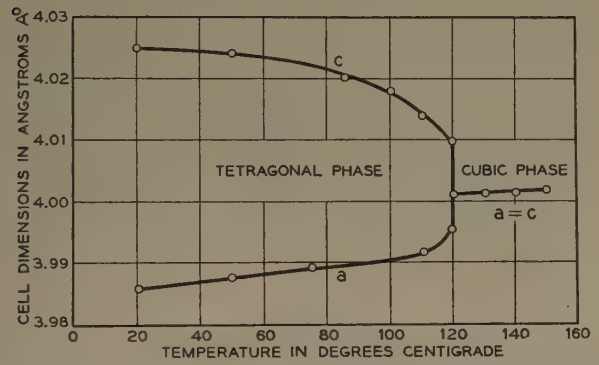

Fig. 3—Cell dimensions of $BaTiO_3$ crystal in cubic and tetragonal phases (after Kay and Voudsen). c=cell dimension along ferroelectric axis; a=cell dimension perpendicular to ferroelectric axis.

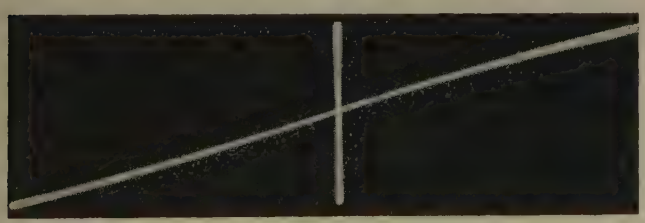

Fig. 4—Nonlinear charge field relation above the Curie temperature (after Merz).

At a temperature slightly above Curie temperature, nearness to the transition causes the polarization field curve to be nonlinear, as indicated by Fig. 4 above. Closer proximity to the Curie temperature causes a larger nonlinear effect. This type of nonlinearity, which can be made use of in dielectric amplifiers, depends rather critically on the temperature difference between the ambient temperature and the Curie temperature and, hence, dielectric amplifiers using this type of nonlinearity are not very useful. Below the Curie temperature, a suitable voltage reverses the direction of the spontaneous polarization and, as shown by Fig. 5, a hysteresis type loop[3] results. When all the cells of the crystal are reversed, the crystal is saturated and gains further polarization only by displacement of the titanium and oxygens toward each other. Since they are very close to each other, the amount of displacement is small and the dielectric constant determined by the slope of the charge-field curve in the saturated region is 200 as compared to a value of 100,000 in the steep part of the curve. The curve of Fig. 5 is for a crystal which initially has all of its unit cells polarized along the thin dimension, and the corners of the hysteresis loop become very sharp. Due to the fast change in polarization with time in the steep section, this part of the loop is hard to photograph. The coercive field E_c, i.e., the field occurring on the zero-charge axis, at the point of charge changeover, is considerably lower than for a multidomain crystal or ceramic and is in the order of 1,000 to 2,500v per cm.

Fig. 5—Hysteresis loop for a single domain crystal (after Merz).

For a square loop type of crystal, the remanent polarization can be measured by measuring the intercept on the polarization curve for zero field. When this is multiplied by the area, the polarization in micro-coulombs per square centimeter is obtained. Results for a particular crystal are shown[3] by Fig. 6 (next page) as a function of the temperature. For this crystal, the polarization rises to about 26 micro-coulombs per square cm at 0 degrees C. At this point, known as the second transition temperature, polarization along the c axis drops by a factor $1/\sqrt{2}$. This is the result of polarization changing from a direction along a [001] axis to a direction along a [011] axis. A third transition temperature occurs at about −80 degrees C. At this temperature the polarization becomes directed along a [111] axis. For this case the polarization along the original direction of measurement is $1/\sqrt{3}$ times that in the 0 degrees to 120 degrees C. range.

[1] H. D. Megaw, "Aging of ferroelectricity in barium titanate and other perovskite-type crystals," *Acta Crystallographica*, vol. 5, pp. 739–749; 1952.
H. T. Evans, "The crystal structure of tetragonal barium titanate," *Acta Crystallographica*, vol. 4, p. 377; 1951.

[2] H. F. Kay and P. Voudsen, "Symmetry changes in barium titanate at low temperatures and their relation to its ferroelectric properties," *Phil. Mag.*, vol. 4, p. 1019; 1949.

[3] W. J. Merz, "Double hysteresis loop of $BaTiO_3$ at the Curie points," *Phys. Rev.*, vol. 91, pp. 513–517; August 1, 1953.

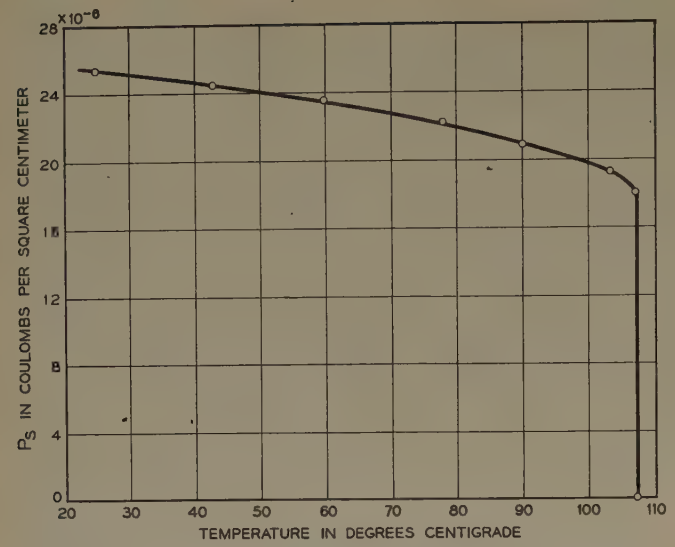

Fig. 6—Spontaneous polarization in first transition range as a function of temperature (after Merz).

These changes in polarization direction are also reflected in changes in the crystal structure.[4] As shown by Fig. 7, the structure becomes tetragonal in the region from 10 degrees to 120 degrees C., orthorhombic from −80 degrees C. to +10 degrees C., and trigonal from

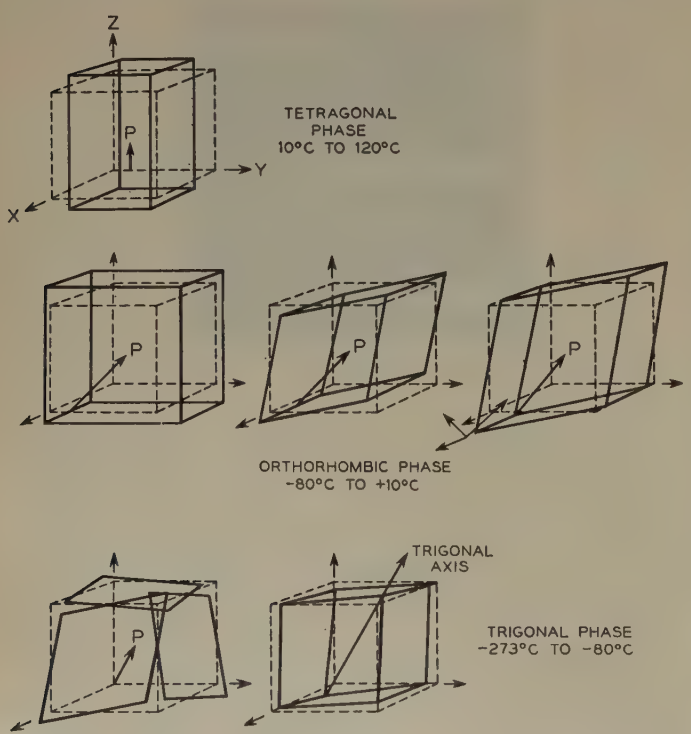

Fig. 7—Spontaneous distortion in the three transition ranges leading to tetragonal, orthorhombic, and trigonal structures.

absolute zero to −80 degrees C. An alternate way of looking at the phase change is to note that the crystal is essentially cubic, but that due to the spontaneous polarization developed in the ferroelectric regions, displacements proportional to squares of the polarization occur, giving the crystals the symmetries observed. The displacements are of the right form to agree with this interpretation[5] and the measured deviations from the cubic form can be used to evaluate the electrostrictive (square law) constants.

Another property of the single crystal that has been measured is the small amplitude dielectric constant. Fig. 8 shows[6] the measured dielectric constant for the *a* axis and *c* axis of the tetragonal state, i.e., perpendicular and parallel to the ferroelectric axis, and some of the constants for the other two regions.

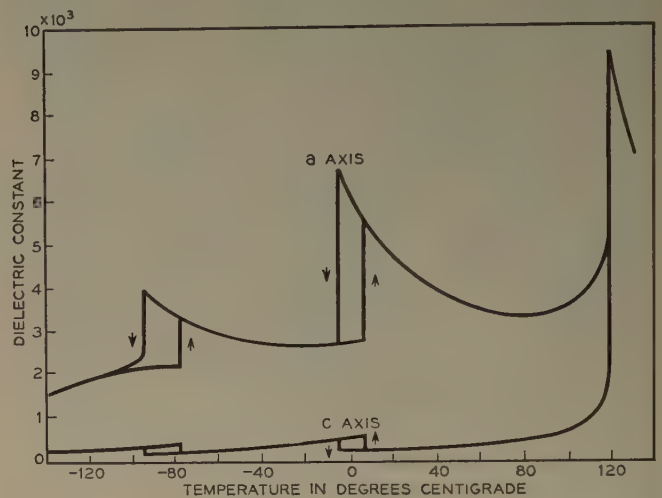

Fig. 8—Small signal dielectric constants of a single domain barium titanate crystal (after Merz).

All properties discussed have to do with the unit cell or single domain crystal. Single crystals can form[7] with a number of domains. The most common twinning type is one in which the *c* axis of one domain is at 90 degrees with the *c* axis of another domain. As shown by Fig. 9 on the next page, these types of domains are separated by walls running nearly 45 degrees with the surface (90-degree walls). In addition, Merz[8] has recently shown that there are 180-degree walls in all single domain crystals. These appear as shown by Fig. 9, when the direction of polarization is being reversed and their quick appearance coincides with the sharp knee of the charge-field hysteresis curve of Fig. 5. It appears to be energetically easier to form new domains than to move 180 degree walls, and the new domains propagate with a finite velocity.

Some recent measurements[9] provide data on the formation time and on the velocity of this domain motion and its variation with temperature.

[4] P. W. Forsberg, "Domain structure and phase transitions in barium titanate," *Phys. Rev.*, vol. 76, p. 1040; 1949.

[5] For a thermodynamic derivation of these terms, see A. F. Devonshire, "Theory of ferroelectrics," *Phil. Mag. Suppl.*, vol. 3, pp. 85–130; April, 1954.

[6] W. J. Merz, "The electric and optical behavior of $BaTiO_3$ single domain crystals," *Phys. Rev.*, vol. 76, pp. 1221–1225; 1949.

[7] For a discussion of twinning in barium titanate, see Forsberg, loc. cit.

[8] W. J. Merz, "Domain properties of $BaTiO_3$," *Phys. Rev.*, vol. 88, pp. 421–422; October 15, 1952.

[9] W. J. Merz, "Domain formation and domain wall motions in ferroelectric $BaTiO_3$ single crystals," *Phys. Rev.*, vol. 95, pp. 690–699; August 1, 1954.

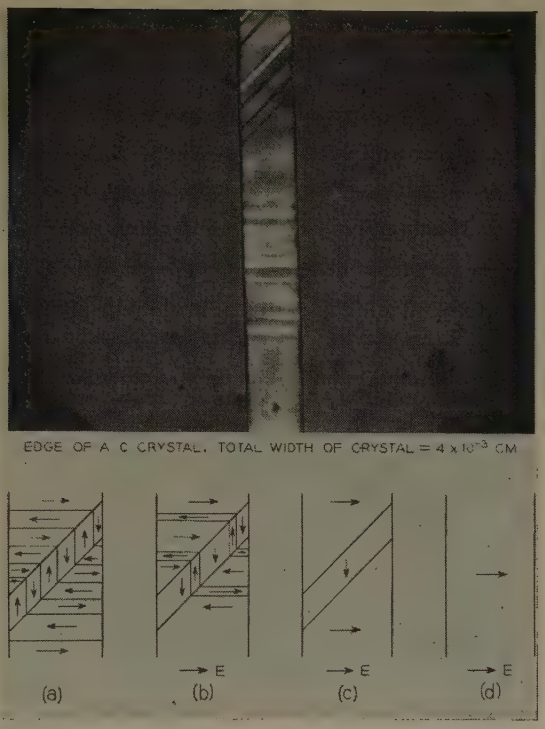

Fig. 9—Photograph of 90-degree and 180-degree domains and relative directions of polarization (after Merz).

Fig. 10 shows the switching-time measured for a crystal 0.005 cm thick. Above a certain field the inverse of the switching-time is proportional to the field. For example, for a temperature of 23 degrees C. and a field of 11,500 volts per centimeter, it takes 0.2 microseconds to reverse the polarization. This corresponds to a velocity of propagation of 2.5×10^4 cm/second. Fig. 10 shows that this velocity increases materially as the temperature rises. These limiting switching-times provide a natural limit to the speed of operation of storage devices and top frequencies for dielectric amplifiers which are in the order of megacycles for thin crystals.

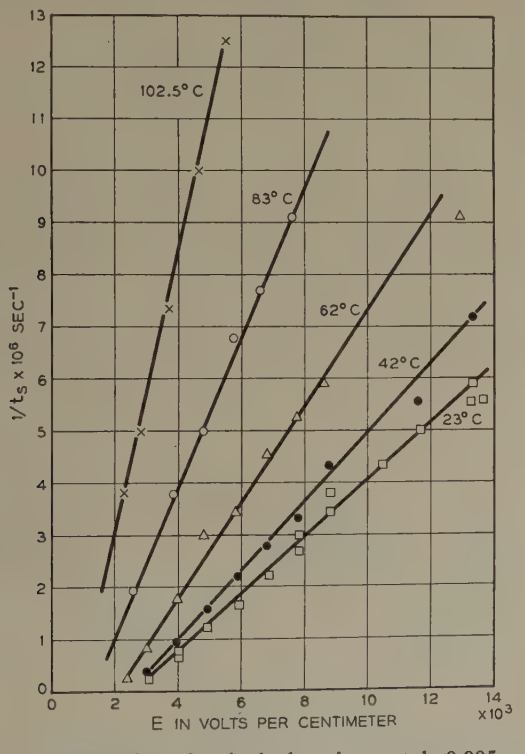

Fig. 10—Switching-times for single domain crystals 0.005 cm thick as a function of the applied voltage (after Merz).

Uses of Barium Titanate Ceramics in Condensers and Transducers

Ceramics can be made from barium titanate and other additants by sintering together crystalline powders of these materials with certain binders to hold them together as they are being sintered. Such ceramics have many of the properties of single crystals. For example, the small signal dielectric constant of a commercial barium titanate ceramic,[10] as shown by Fig. 11, has the same type of dielectric behavior and the same transition temperatures as the single domain crystal. The dielectric constant of the ceramic closely approaches that for a polycrystalline material[11] and it appears close to what one would obtain by assuming that all orientations of the domains are equally probable. The power factor tan δ is between 1 and 2 per cent in the first ferroelectric range but is considerably higher in other ranges.

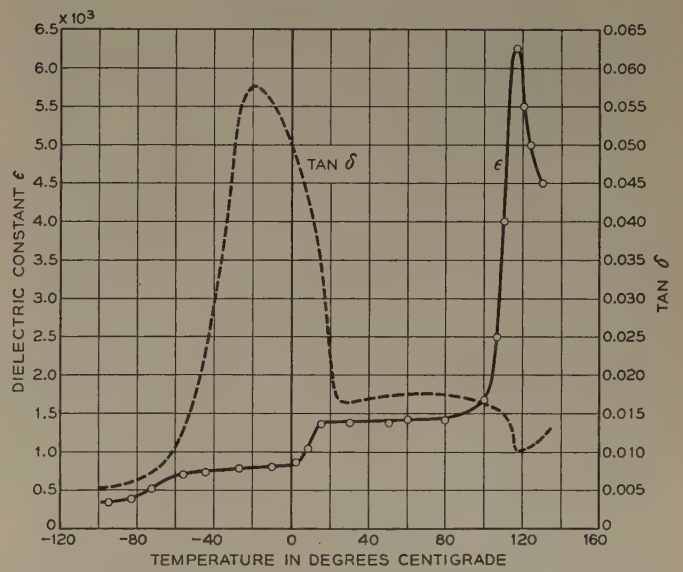

Fig. 11—Dielectric constant and dissipation factor of a polycrystalline barium titanate ceramic (after Von Hippel, Breckenridge, Chesley, and Tisza).

Since the dielectric constant and the power factor in the room temperature range show large variations which are associated with the presence of the second transition temperature, it is desirable to be able to move this transition point to lower temperatures. Two additives have been commonly used for this purpose, lead titanate and

[10] A. Von Hippel, R. G. Breckinridge, F. G. Chesley and L. Tisza, "High dielectric constant ceramics," *Jour. Ind. Eng. Chem.*, vol. 38, p. 1097; 1946.

[11] B. T. Matthias and A. Von Hippel, "Domain structure and dielectric response of barium titanate single crystals," *Phys. Rev.*, vol. 73, pp. 1378–1384; June 1, 1948.

calcium titanate. The lead titanate[12] raises the Curie temperature and lowers the second transition temperature, as shown by Fig. 12, while calcium titanate[13] leaves the Curie temperature unchanged and lowers the second transition temperature. Combinations of lead and calcium titanate additions to barium titanate have also been used, with the results shown by Fig. 22. These have lower dielectric constants but also lower power factors, which are in the order of 0.25 per cent for low field strengths. The use of titanates in capacitors is probably the largest use so far made.

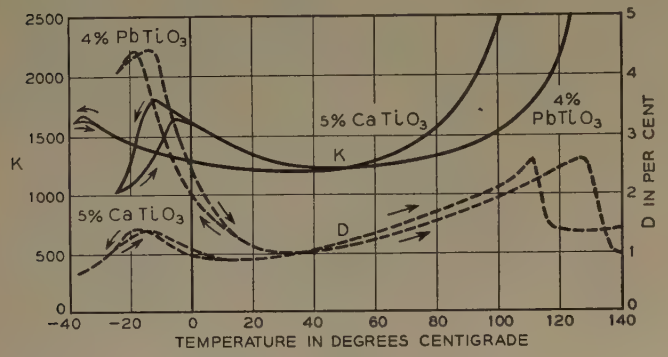

Fig. 12—Dielectric constants and dissipation factors for titanate ceramics with calcium and lead titanate additions (after Jaffe).

Another large and growing use of titanates is their application in electromechanical transducers. This effect arises from the fact that the single crystal, as shown by Fig. 3, increases in dimension in the direction of spontaneous polarization. In the ceramic, all directions for the ferroelectric axis are equally probable and hence the ceramic appears unpolarized. However, as shown by Fig. 13(a), if one applies a field normal to the thickness direction, the directions of the ferroelectric axis of a number of domains are changed to be more nearly parallel to the thickness, by the titaniums changing position from one oxygen to another more nearly in the direction of the applied field. Stress-charge measurements[14] indicate that about one-third of the total number of domains can be lined up by this process, and since the effect is a square-law one, expansions about 1/9 of that for a single crystal result, giving increases in thickness of about 700 parts in 10^6. This increase is considerably larger than that for most magnetostrictive materials. In addition to the increase in thickness, a decrease occurs in the radial direction a little less than half as large as the thickness increase. This mode can be used to drive a radial resonance in a disc of the material or a length vibration in a slab, as shown by Fig. 13(b). A third mode occurs if one polarizes the ceramic in one direction and applies a driving force at right angles. This combination produces a shearing mode as shown by the dashed line of Fig. 13(c).

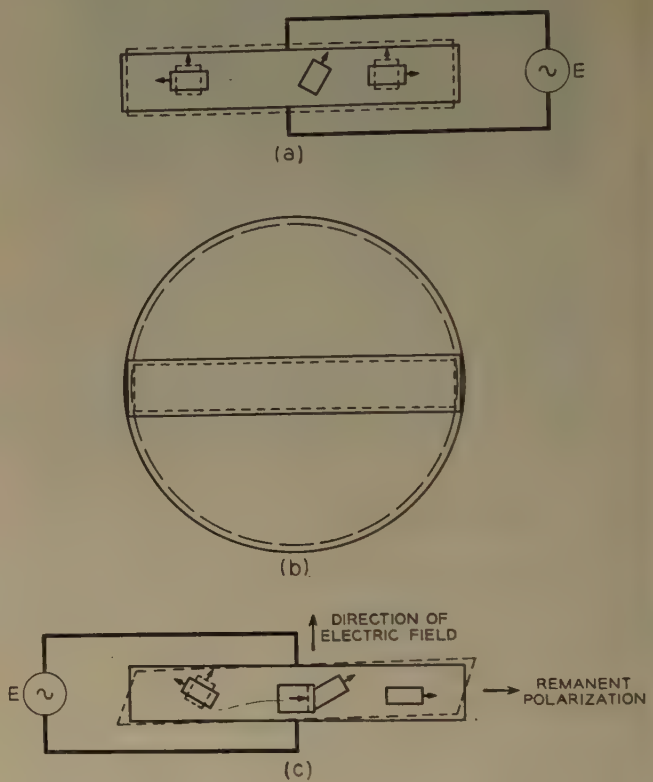

Fig. 13—Distortions produced by poling of titanate ceramics.

As is obvious from the data of Fig. 3 and Fig. 7, a change in the unit cell is proportional to the square of the electric displacement. Hence, if one plots the thickness strain against the applied field, a butterfly loop results as shown by Fig. 14. The top curve shows the electric displacement as a function of field strength and the minima in the strain occur when the electric displacement goes through zero. If now one puts a high polarizing voltage on the ceramic and removes it, the ceramic will have a remanent polarization, as seen from the top figure of Fig. 14, and hence the strain will have a finite value. The stress-field relation will be the curve on the top butterfly loop and hence a field applied in the direction of polarization will cause an increase in thickness while the reverse field causes a decrease in thickness. Hence, an effective piezoelectric constant results and the polarized ceramic acts as a piezoelectric crystal.

One very useful application of the polarized ceramic is in measuring small or large dynamic forces. Fig. 15 shows the two types of units used. The one on the left is poled in the direction of the thickness and responds to a normal force. The sensitivity depends on the distribu-

[12] W. P. Mason, "Piezoelectric Crystals and Their Application to Ultrasonics," D. Van Nostrand Co., Inc., New York, N. Y., chap. XII; 1950.

[13] E. N. Bunting, G. R. Shelton and A. S. Creamer, "Properties of calcium-barium titanate dielectrics," *Jour. Res. Nat. Bur. Stand.*, vol. 43, pp. 237–244; 1949. H. Jaffe, Brush Stokes, Cleveland, Ohio; December, 1951.

[14] W. P. Mason, "Use of temperature and time stabilized barium titanate ceramics," Proc. First ICA Cong. Electro-Acoustics, W. C. Meinema Ltd.; *Delft and Acustica*, vol. 4, pp. 200–202; January, 1954.

tion of stress over the surface. For a point application of force, the sensitivity of the unit in volts is equal to

$$V = \frac{0.56 \times 10^{-2} Ft}{A} \tag{1}$$

where F is the applied force in grams, t the thickness in centimeters, and A the cross-sectional area in square centimeters. The sensitivity, which is about 50 times that

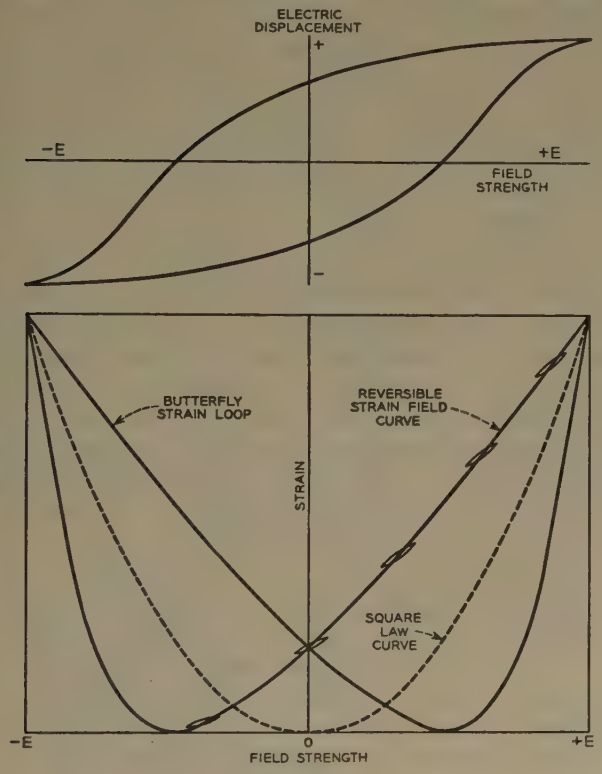

Fig. 14—Butterfly strain loop and its relation to electric displacement-electric field hysteresis loop.

for quartz, is sufficient to give a response to forces as low as 1 gram when the unit is connected to a standard type oscilloscope. To measure forces lasting relatively long times, the unit has to be connected to an input amplifier having a high leakage resistance, since otherwise the

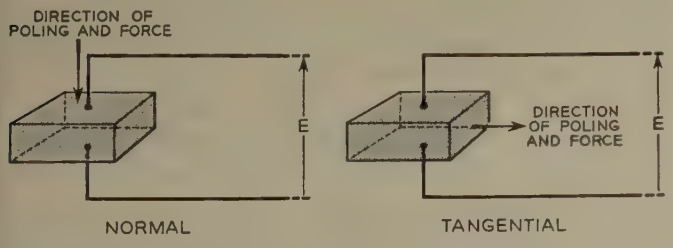

Fig. 15—Titanate ceramic units responding to normal and tangential forces.

charge generated by the applied force will leak off in too short a time. The generated charge leaks off according to the equation

$$Q = Q_0 e^{-t/RC}. \tag{2}$$

Hence, for a small unit 0.05 cm thick by 0.01 square cms having a capacitance of 20 $\mu\mu$f, as has been used for relay force measurements,[15] we require an amplifier with an input impedance of 750 megohms to measure a force lasting 0.015 seconds. Fig. 16 shows the circuit of such an amplifier using a 6SU7 double triode with feedback between stages to obtain the high input impedance. On account of the high impedance, this tube is sensitive to 60-cycle pickup, and this has to be balanced out by using the 60-cycle input circuit. Such units have been used to measure dynamic relay forces[15] for intervals of time from 0.015 seconds down to microseconds. For the unit described above, the lowest frequency resonances are in the order of 2 megacycles and hence the response will be a replica of the applied force up to times in the order of 1 microsecond.

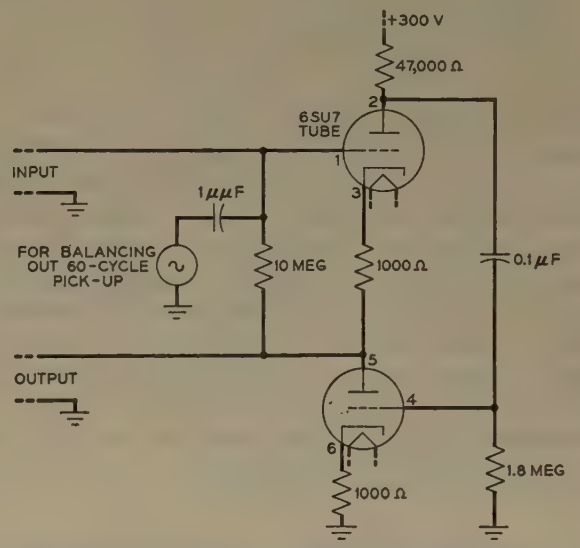

Fig. 16—High input impedance double triode vacuum tube circuit for measuring voltages corresponding to slowly impressed forces (after Head).

The tangential force measuring crystal of Fig. 15 has a sensitivity independent of the stress distribution and equal to

$$V = \frac{1.94 \times 10^{-2} tF}{A} \tag{3}$$

where V is in volts and F in grams. One use that this type of force-measuring ceramic has been put to is in studying the forces occurring in frictional wear. For this purpose, a barium titanate transducer and transforming mechanical horn,[16] capable of giving a 0.002-inch displacement at 18,000 cycles, is used to produce the wear. This moves tangentially on the surface of a thin piece of material whose wear properties are to be

[15] W. P. Mason and S. D. White, "New techniques for measuring forces and wear in telephone switching apparatus," *Bell Sys. Tech. Jour.*, vol. 31, pp. 469–503; May, 1952.

[16] W. P. Mason and R. F. Wick, "A barium titanate transducer capable of large motion at an ultrasonic frequency," *Jour. Acous. Soc. Amer.*, vol. 23, pp. 209–214; March, 1951.

studied. The material is in turn cemented to a tangential force-responding ceramic and to the arm of a pivot which is used to impress a definite static force against the wire. The arrangement is shown by Fig. 17. The tangential force applied to the material is shown by the voltage generated by the ceramic. Fig. 18 shows the traces of such voltages for a number of different amplitudes of motion. For bakelite plastics, when the amplitude of motion is less than 0.05 mil inches, a sinusoidal voltage is obtained and corresponding wear measurements show very little effect. This region corresponds to a region where the points of contact bend elastically without slipping on each other. As the amplitude of the motion is increased, a value is reached for which the points of contact can no longer bend sufficiently and slip occurs during which some points lose contact and other points form contacts. That this is a stick-slip process is shown by the fact that for amplitudes of 0.5 mil inches, for example, the same pattern repeats itself for a time of the order of seconds to minutes. Since the wire is moving at a frequency of 18,000 cycles per second, this indicates that wear occurs by the bending of the same high points for many thousands of times until they finally fatigue and break off. At larger amplitudes the pattern changes much faster. Wear measurements indicate that the wear increases nearly linearly with the amplitude and also with static force. However, there is a finite amplitude region corresponding to the region of elastic response, where no measurable wear occurs. According to calculation,[17] the limiting amplitude for elastic response δ_x is related to the elastic constants, applied static force N and radius of the contact sphere, by the equation

$$\delta_x = \frac{\left(\frac{3N}{\mu}\right)^{2/3}(2-\sigma)f}{8\sqrt{r(1-\sigma)}} \qquad (4)$$

where μ is the shearing modulus, σ is Poisson's ratio and f the coefficient of friction.

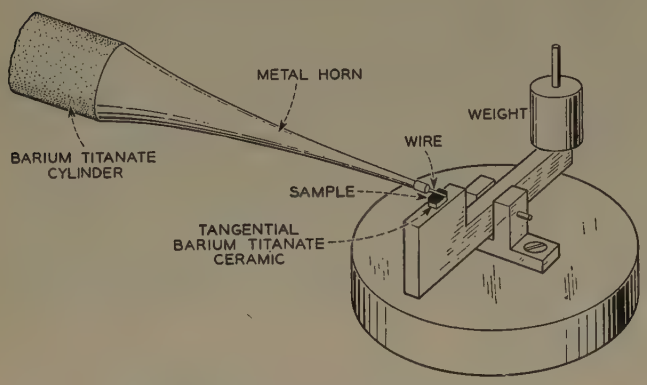

Fig. 17—Titanate ceramic and exponential metal horn for producing wear in materials.

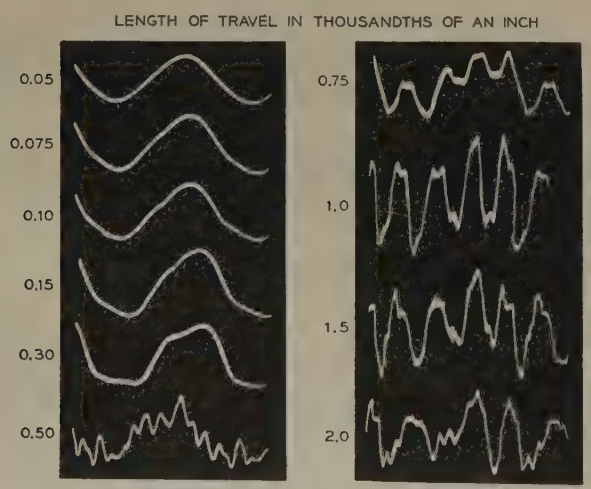

Fig. 18—Force responses for various travel lengths.

Many other uses have been found for barium titanate transducers. On account of the fact that they can be sintered in any desired form, they have been made in the form of focusing and diverging sound radiators. Large numbers of plates and rings are used in underwater sound applications. About 15 per cent of all phonograph pickups are made using barium titanate ceramics. Another application is in sensitive accelerometers. Fig. 19 shows two types[18] of accelerometers that have been used in vacuum tube testing. The one on the right has a central electrode on the inside surface and two electrodes on the outer surface, as shown. If the two outside electrodes are connected in parallel, the accelerometer measures the acceleration along the

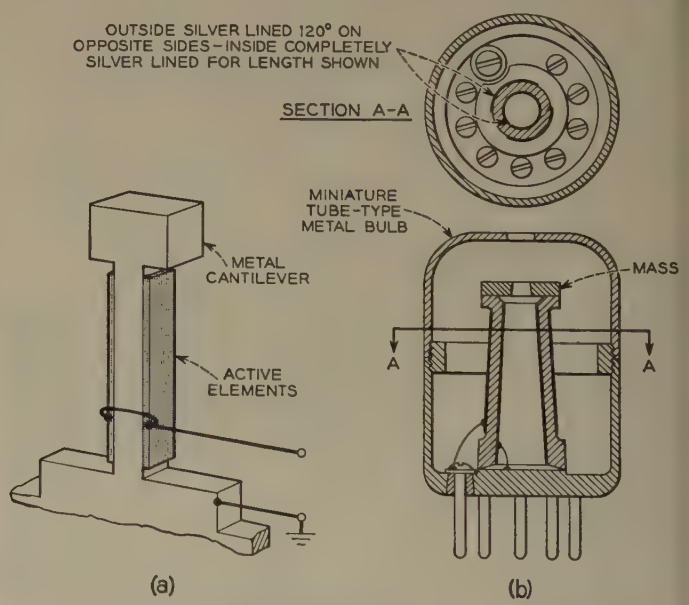

Fig. 19—Two types of barium titanate accelerometers (after Stubner).

[17] R. D. Mindlin, W. P. Mason, T. F. Osmer, and H. Deresiewicz, "Effects of an oscillating tangential force on the contact surfaces of elastic spheres," *Proc. First Natl. Cong. Appl. Mech.*, pp. 203–208; 1952.

[18] F. W. Stubner, "Acceleration effects on electron tubes," *Bell Sys. Tech. Jour.*, vol. XXXII, pp. 1203–1231; September, 1953.

length of the tube. If the voltage is taken off the two outside electrodes, the acceleration is measured normal to the center line of the electrodes.

In all the transducer applications of barium titanate, it is desirable that the coefficients of the transducer should remain constant with temperature. This is particularly true for force and acceleration measurements where a definite calibration is required which should not vary too much with temperature or show temperature hysteresis. Furthermore, other applications, such as electromechanical filters and delay lines, could be made if the stability were considerably greater and if the mechanical and electrical losses in the ceramic could be greatly lowered. Hence, a study of these possibilities has recently been made.[14] Fig. 20 shows measurements as a

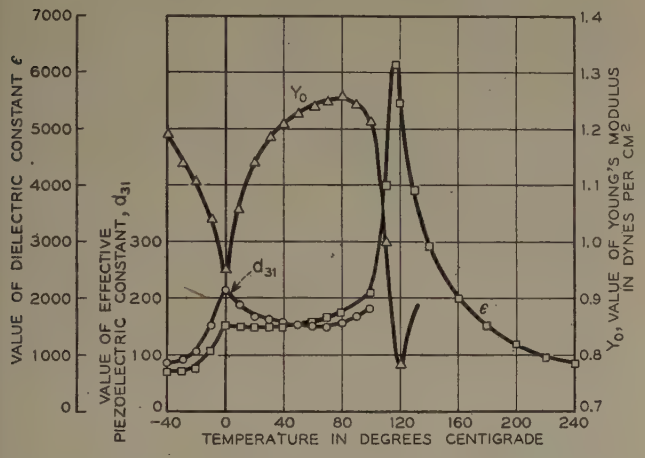

Fig. 20—Dielectric constant, effective piezoelectric constant and Young's modulus as a function of temperature for a commercial barium titanate ceramic.

function of the three parameters of interest in a radial mode transducer,—the value of Young's modulus Y_0, the radial mode piezoelectric constant d_{31}, and the dielectric constant,—and it is seen that very large variations occur in all these parameters for barium titanate ceramics, particularly near the transition temperatures. In addition, the ratios of reactance to resistance, i.e., the Q's for mechanical and electrical quantities, are only about 200 and 75 respectively. It was pointed out previously that the dielectric constant could be stabilized by separating the transition temperatures with the use of additions such as lead and calcium titanate. Other studies[19] have shown that considerable improvement can also be made in the other parameters by similar techniques. But, even using large amounts of each addition separately does not result in ceramics stable enough for such applications as filters and delay lines.

By combining[14] fairly large percentages of both lead titanate and calcium titanate, considerably more stable ceramics can be obtained. Fig. 21 shows the value of Young's modulus for various compositions measured from −40 degrees C. to +100 degrees C. It is obvious that considerable improvement is effected by increasing the lead and calcium content. For compositions having 8 per cent or more by weight of both additants, the value of Young's modulus peaks at 25 degrees C., and from 13 degrees to 40 degrees C. (the usual indoor temperature range) the modulus does not vary by more than 0.1 per cent. The composition 8.3 per cent of $CaTiO_3$, 12 per cent $PbTiO_3$, 79.7 per cent $BaTiO_3$, does not vary in modulus by more than 0.8 per cent from −40 de-

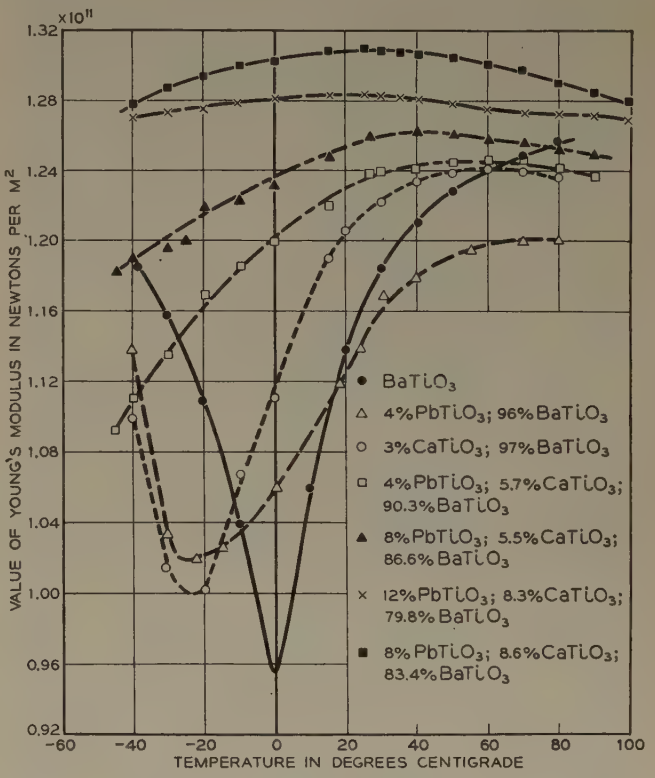

Fig. 21—Effect of lead and calcium titanate additives on the value of Young's modulus.

grees C. to +80 degrees C. Fig. 22 (next page) shows the corresponding dielectric and piezoelectric constants and, there also, the variations are much reduced. The absolute values of both the dielectric and piezoelectric constants decrease very materially. However, the electromechanical coupling factor k, which, for a radial mode, is given by the equation

$$k = d_{31}\sqrt{\frac{8\pi Y_0}{\epsilon(1-\sigma)}} \qquad (5)$$

where σ is the value of Poisson's ratio, has fallen off only 25 per cent for the lowest coupled material. Furthermore, the mechanical and electrical Q's have increased by factors of 4 initially. These factors increase with aging, as discussed next, so that mechanical Q's of 1,500 to 2,000 are possible with these compositions. Hence, the constants are sufficiently good to be considered for applications in filters and delay lines.

[19] Mason, op. cit.
D. A. Balincourt and F. Kulcsar, "Electromechanical properties of $BaTiO_3$ compositions showing substantial shifts in phase transition points," *Jour. Acous. Soc. Amer.*, vol. 24, pp. 709–714; November, 1952.

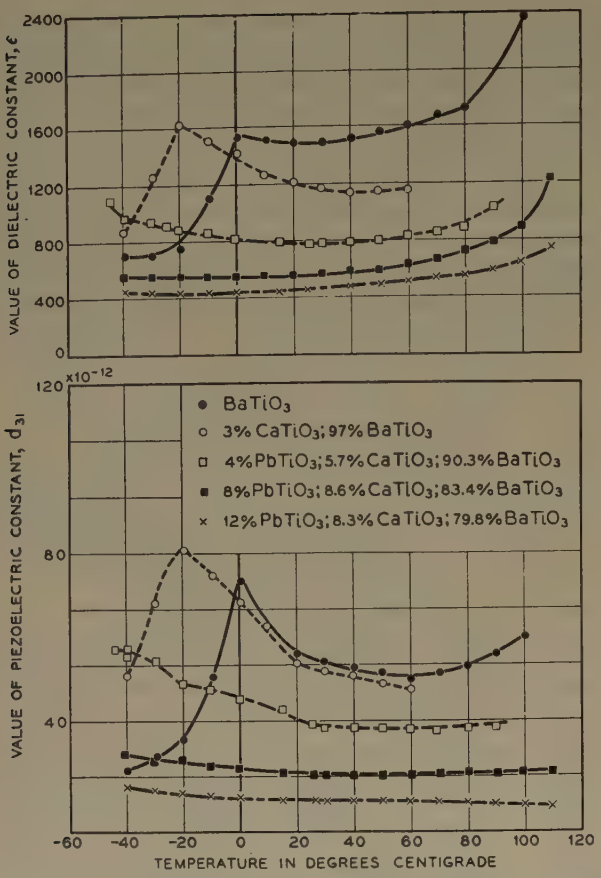

Fig. 22—Effect of lead and calcium titanate additions on the dielectric constants and effective piezoelectric constants of barium titanate ceramics.

Another source of instability in barium titanate ceramics is a long time-aging effect. Fig. 23 shows the increase in frequency and decrease in coupling factor for the two most stable ceramics. The frequencies increase by 1.5 to 2 per cent in the course of a year and are still aging, at a much lower rate, out to 2 years. The electromechanical coupling factor decreases by about 15 per cent during this time and appears to be stabilized at the end of that time. Fig. 24 shows the other two parameters of interest, the dielectric constant and the Q as a function of time at 25 degrees C. The Q more than doubles in the course of a year and approaches values of 1,600 to 2,000, while the dielectric constant decreases about 20 per cent during this time. All these values are for polarized ceramics, but it is well known that the dielectric constant changes by similar amounts, and in the same time range, for the unpolarized ceramics also. In order to see whether this time effect was connected with a process having an activation energy,

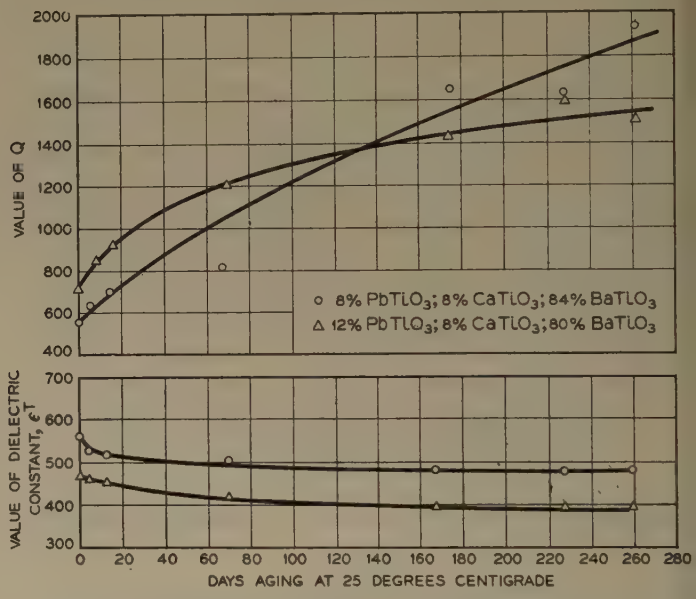

Fig. 24—Aging effects on the Q and dielectric constant of two ceramic mixtures at 25 degrees C.

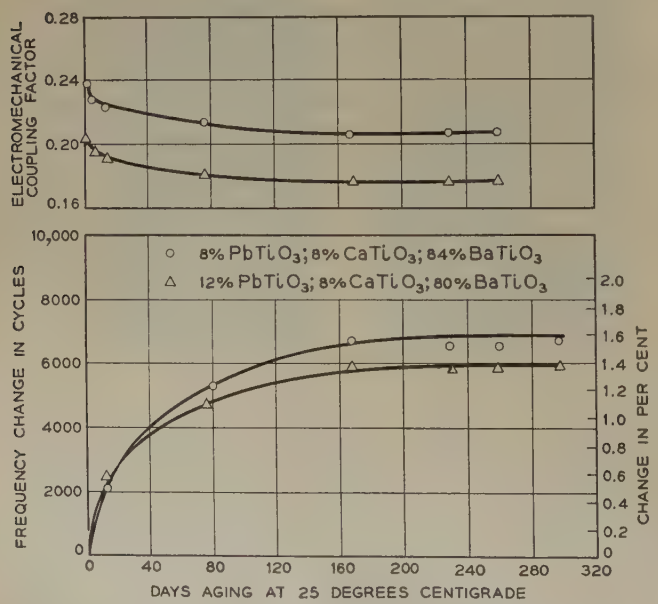

Fig. 23—Aging effects on the frequency and electromechanical coupling factors of two ceramic mixtures at 25 degrees C.

another set of aging curves were run at 70 degrees C. The temperature has to be under the Curie temperature of 120 degrees C., for otherwise the ceramic will be in the cubic state. It has been found too, that the initial conditions are repeated and the aging starts all over again when the ceramic is repoled by heating it above the Curie temperature and cooling under a field. The aging curves for 70 degrees C. are shown by Fig. 25 for the change in frequency, the coupling, and the dielectric constant. All these aging effects occur much more rapidly than at room temperature. If we use the usual equation for determining the activation energy,

$$\log_e\left(\frac{t_2}{t_1}\right)\frac{RT_1T_2}{T_2 - T_1} = H \qquad (6)$$

where t_2 and t_1 are the times required to complete a given portion of the entire aging cycle at the absolute temperatures T_2 and T_1, and R is the gas constant (2 calories per mole), H, given in calories per mole, is about 19,000.

The high activation energy found indicates that a heat treatment can be developed which will remove a large part of the variation due to aging. The best cycle found for aging is to heat the ceramic to 75 degrees C. for two weeks, then to reduce the temperature to 45 degrees C. for another two weeks. After this treatment, the dielectric constant and electromechanical coupling do not change measurably. The frequency still increases slightly, but over a year's time the increase is less than 0.2 per cent and appears stabilized at the end of this time. This is a small enough variation to be allowed for in the design of a filter or delay line.

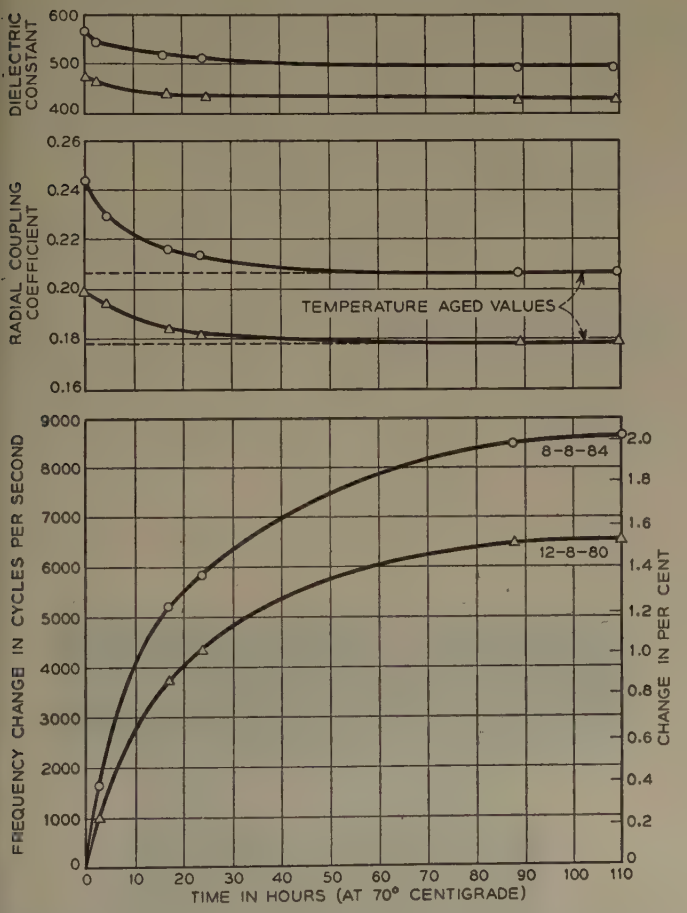

Fig. 25—Aging effects on the dielectric constant, radial electromechanical coupling factor and resonant frequency at 70 degrees C.

Use of Nonlinear Effects in Barium Titanate in Dielectric Amplifiers and Storage Devices

In addition to the linear dielectric effects and linear polarized electromechanical effects discussed in the above section, several nonlinear effects are under investigation and may eventually be widely used. These effects find applications in frequency modulated oscillators, in information storage devices and in dielectric amplifiers.

The simplest of these devices, the biased frequency-modulation oscillator depends on the fact that the capacitance is a function of a dc or low frequency bias. For a barium titanate ceramic, the relation between the bias and the dielectric constant is shown by Fig. 26. With a bias of 30 kv per cm the dielectric constant is reduced to one-half. If one limits the bias to 14,000v per cm, a relatively straight-line relation between dielectric constant and field is obtained. For a thickness of 0.005 inch of ceramic, a biasing voltage of 70v, and a signalling voltage of ± 70v, a frequency change of ± 7.5 per cent is obtained which is sufficient for most frequency modulated oscillators.

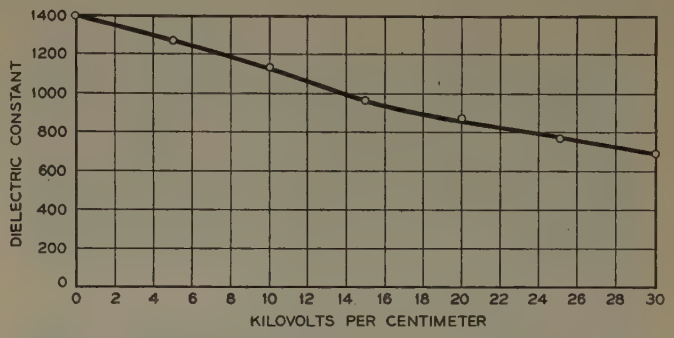

Fig. 26—Decrease in small signal dielectric constant of a barium titanate ceramic due an applied dc field (after Danielson).

Probably the widest potential use of nonlinear effects in ferroelectrics is in information storage. This makes use of the square-shaped hysteresis loop that can be obtained for a single crystal as shown by Fig. 5. Positive and negative pulses corresponding to the numbers 1 and 0 in the binary system can be stored and the information can be read out by the difference in response to a probing pulse. The advantages and possible applications of this system have been described in detail by Anderson.[20]

Another potentially important device is the dielectric amplifier which uses the nonlinear characteristics of the charge-field curves. So far this a research product which has promising properties. Some types suffer from temperature instability but by using single crystals below the Curie temperature, as discussed later, this instability can be reduced. The frequency range for which the dielectric amplifier can be used is limited but it appears to be adequate for many carrier and perhaps low frequency radio applications. Preliminary tests on the aging of hysteresis loops are encouraging, but in the absence of long time-aging tests of complete amplifiers, aging properties of the device represent one of the open questions to be settled before applications are made.

The dielectric amplifier is the analog of the well known magnetic amplifier. The principle of this device is illustrated by Fig. 27 on the next page, although this is only one of many usable circuits. A carrier frequency f_c is connected in series with the two ferroelectric condensers and a load resistance R_L, and if we adjust the

[20] J. R. Anderson, "Ferroelectric storage elements for digital computers and switching systems," *Elec. Eng.*, vol. 71, pp. 916–922; October, 1952.

load resistance to match the impedance of the condensers at the carrier frequency, the carrier power output is equal to

$$P_C = I^2R = (2\pi f_c Q_0)^2 \times \frac{1}{\frac{2\pi f_c C_0}{2}} = \frac{4\pi f_c Q_0^2}{C_0} \qquad (7)$$

where Q_0 is the amplitude of the charge variation. The circuit on the left consists of two inductances L_0 which are large enough to prevent the carrier current from traversing this path, a biasing battery, and a low frequency signal source having an internal resistance R_S.

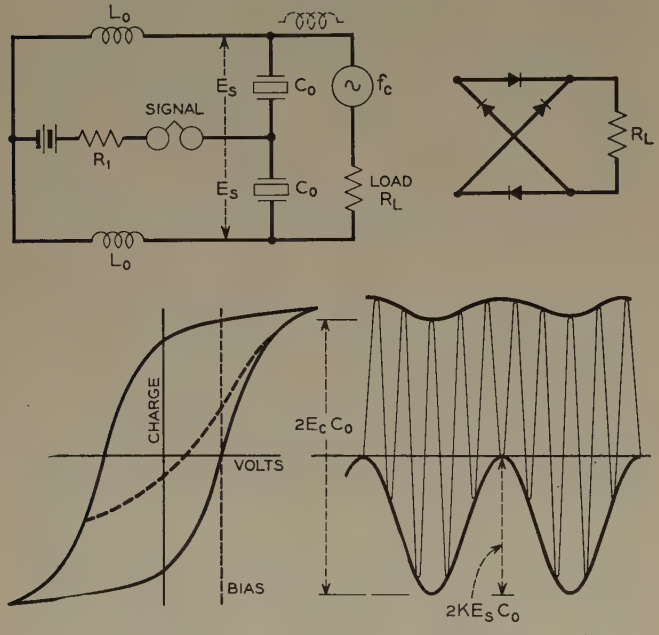

Fig. 27—Circuit for a dielectric amplifier.

The effect of the biasing field is to cause the loop traversed to change from the solid line to the dotted line. Some actual oscillograph patterns of the effect of a bias are shown by the photographs of Fig. 28. These are for a barium titanate single crystal 0.004 inches thick with a peak ac voltage of 28v. The lower figures are for a 9v bias and 13.5v bias. Hence, it is evident that a bias cuts down the amount of charge flowing through the capacitances C_0 and, hence, if a sinusoidal signal is impressed on the ferroelectric condensers C_0, the output is modulated as shown by the top right figure. The static bias is necessary in order that the modulation will be at the frequency of the impressed signal; otherwise the modulation will be at twice the frequency. The envelope of modulation will be proportional to the signal E_S or

$$Q = \frac{C_0}{2} KE_S \qquad (8)$$

where K is the voltage modulation factor. For the circuit without tuning inductance, this factor cannot be larger than two. If now we rectify the modulated output by using the rectifying circuit shown, the rectified power output will be

$$\text{P.O.} = \pi f_c C_0 K^2 E_S^2. \qquad (9)$$

To calculate the power input, we note that at the low signal frequency the load can be regarded as a short circuit and, hence, the input circuit will be the load R_1 in series with the inductance $L_0/2$ and the capacitance $2C_0$. Since the signal voltage E_S is always across the condenser $2C_0$, the power input is

$$\text{P.I.} = i_S^2 R_S = [4\pi f_S C_0 E_S]^2 R_S. \qquad (10)$$

Since the inductance $L_0/2$ and capacitance $2C_0$ are in the form of low-pass filter with an input impedance $Z_0 = 1/4\pi f_0 C_0$, where f_0 is the upper cutoff, the best value for R_S is Z_0. Hence, the power input is equal to

$$\text{P.I.} = 4\pi f_S^2 E_S^2 C_0 / f_0 \qquad (11)$$

and the power gain of the amplifier is

$$\text{P.G.} = \frac{\text{P.O.}}{\text{P.I.}} = \frac{\pi f_c C_0 K^2 E_S^2 f_0}{4\pi f_S^2 E_S^2 C_0} = \frac{K^2}{4} \frac{f_0}{f_S} \frac{f_c}{f_S}. \qquad (12)$$

The most critical range is the upper frequency region where $f_S \sim f_0$ and hence effectively the gain is $(K^2/4) f_c/f_0$. Some further gain can be obtained by tuning the carrier frequency by the coil shown by the dotted lines. This can be connected either in series or in shunt. It has the effect of introducing a tuning factor

$$\text{T.F.} = \left(\frac{1 + AX}{1 - BX + CX^2}\right)^2 \qquad (13)$$

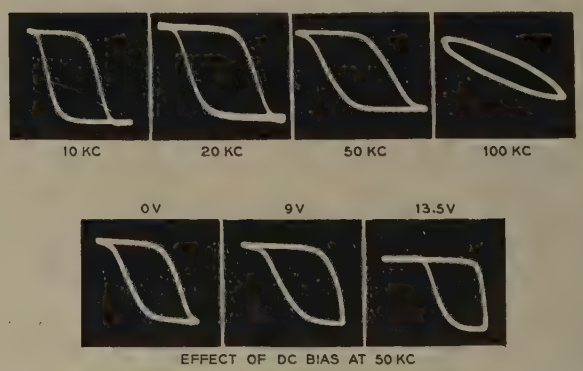

Fig. 28—Hysteresis loops for a crystal 0.01 cm thick as a function of frequency and bias.

where A, B and C are factors discussed in the Appendix, and $X = 1 - (f_t/f_c)^2$, where f_t is the tuning frequency $1/2\pi\sqrt{L_0 C_0/2}$. If the tuning frequency is less than the carrier frequency, f_c the factor X is positive. On account of the negative term in B, the denominator can be made to approach zero and a higher gain obtained, at the expense of a narrower bandwidth. This in effect produces a positive feedback and instability may set in if the effect is made too large. If, however, the tuning frequency is stable enough with temperature, some gain may be obtained by using this factor.

So far most of the work on dielectric amplifiers[21] has been done above the Curie temperature, where, as shown by Fig. 4, the nonlinearity is not accompanied by appreciable hysteresis effects. Although reasonable gains are obtained, the nonlinearity is very critical with temperature and thus this method does not result in a very useful amplifier. Furthermore, since the dielectric constant changes so rapidly with temperature, no use can be made of the tuning factor without courting instability. All magnetic amplifiers are used in regions much below the Curie temperature, where the flux-magnetic field curves are stable with temperature and hence a study has been made of the possibility of using ferroelectrics in this region. One obvious fact is that a single crystal will give much better results than the ceramic on account of its much narrower hysteresis loop and the much sharper change of slope with field. Heating of the crystal is one source of difficulty with the use of ferroelectrics below the Curie temperature. This is indicated by the hysteresis loops shown by Fig. 28 which are for a crystal 0.004-inch thick with a 0.020-inch diameter electrode mounted in air with a peak voltage of 28v applied. Above 50 kc, the temperature has risen above the Curie temperature and the nonlinearity has practically disappeared. With crystals of this sort, amplifiers have been constructed which give gains of 10 db over a band of 3,000 cycles with a carrier frequency of 50 kc.

Recently, however, a few experimental crystals have been made[22] in thicknesses of 0.002-inch which are saturated by voltages of 10v. Furthermore, with electrodes 0.004 inch or less in width, the heating effect is so small that the hysteresis loops remain unchanged up to 200 kc or higher. Fig. 29 shows the hysteresis loops

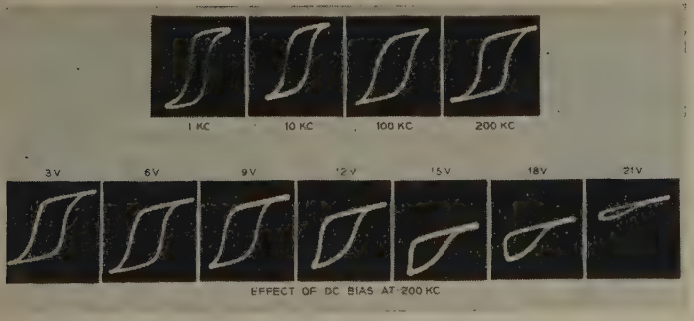

Fig. 29—Hysteresis loops for a crystal 0.005 cm thick as a function of frequency and bias.

for an rms voltage of 20v. Up to 200 kc the shape of the loop remains unchanged. This was the top frequency of the hysteresis loop viewer and in all probability the loop remains unchanged up to 1,000 kc. Fig. 29 also shows the effect of biases from 3v to 21v on the shape of the hysteresis loop at the carrier frequency of 200 kc. Fig. 30 is an oscillographic sweep picture of the effect of a modulation voltage on the envelope of the carrier output. The sweep is not fast enough to resolve the 200-kc carrier. Two frequencies of modulation are used, 1,500 cycles and 7,000 cycles. From low frequencies up to 7,000 cycles the amount of modulation is independent of the frequency, and as can be seen by comparing the modulation with the rms value of signal voltage, the modulation is quite linear. Above 7,000 cycles the amount of modulation falls off rapidly with frequency.

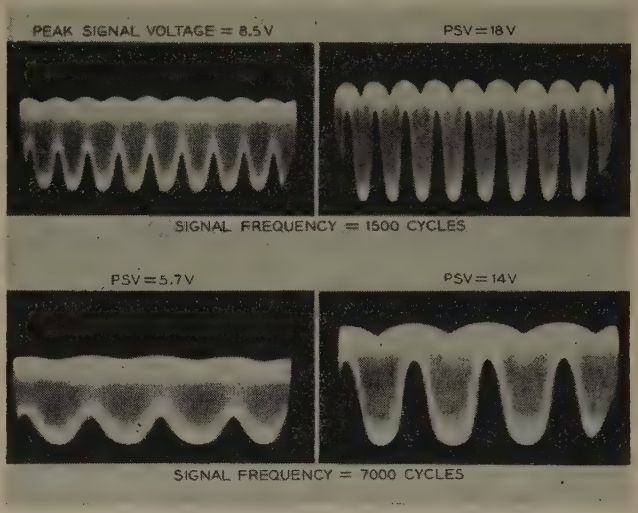

Fig. 30—Modulated carrier output for a dielectric amplifier for two signal frequencies.

This crystal gave a power amplification of about 12 db over a frequency band of 7,000 cycles. With crystals of these types, moderate and high gain dielectric amplifiers are possible. Using single crystals below the Curie temperature, they are not temperature sensitive. Care has to be taken to mechanically damp the crystals or they are likely to be broken by mechanical vibrations.

As shown by Fig. 10, there is a definite limit to the time required to reverse a domain. This time is proportional to the thickness of the crystal and, for a crystal 0.005 cm thick, will limit the upper frequency of operation to a few megacycles even if all heating effects are removed. Hence, it appears that the most probable applications are for carrier systems and low frequency radio system.

The Appendix gives a theoretical study of the factors affecting the gain of a dielectric amplifier assuming simplified hysteresis loop forms which, however, are rather close to the forms measured.

Appendix

Effect of Nonlinearity and Hysteresis on the Equations for a Dielectric Amplifier

The circuit of Fig. 27, with a shunt tuning coil can be reduced in principle to that shown by Fig. 31(a). In this representation the coil L represents the tuning

[21] A. M. Vincent, "Dielectric amplifier fundamentals," *Electronics*, vol. 24, pp. 84–88; December, 1951.
H. Urkowitz, "A ferroelectric amplifier," *Jour. Frank. Inst.*, vol. 254, pp. 517–530; December, 1952.
[22] Supplied by J. R. Anderson.

inductance, C_c and R_c the effective capacitance and resistance of the ferroelectric condenser at the carrier frequency, and R_2 represents the input impedance of the rectifier load. i_c represents the carrier current and V_1 the modulating voltage. This voltage is supplied by the circuit of Fig. 31(b), where R_1 is the load impedance at the signal frequency, e_S the signal voltage, L_0 the input inductance, and C_0 the capacitance of the ferroelectric condenser at the signal frequency. This may be different than C_c due to change in the shape of the hysteresis loop with frequency.

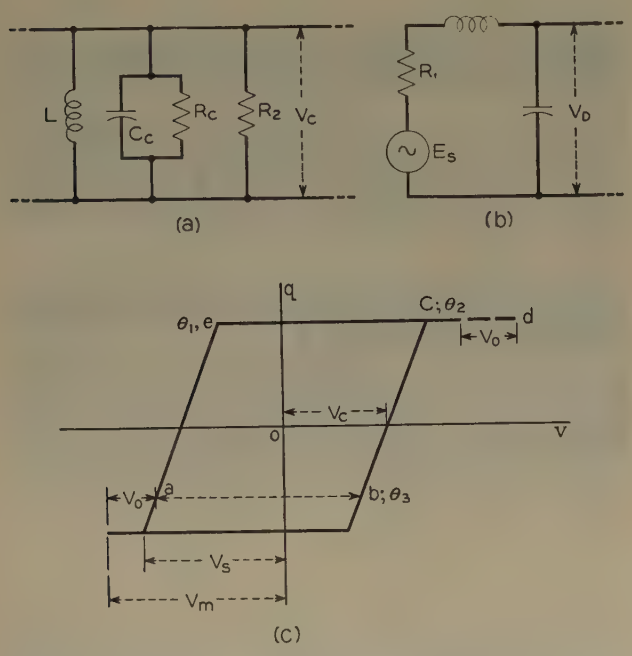

Fig. 31—Equivalent circuit and idealized hysteresis loops of use in calculating gains for dielectric amplifiers.

If V_m is the peak value of the carrier voltage across the circuit of Fig. 31(a), and i_{0_m} the peak value of the input current i_0, which is assumed constant, we can write

$$V_m = i_{0_m} |Z(V_m, V_0)|$$

$$= i_{0_m} \frac{1}{\sqrt{\left(\frac{1}{R_2}+\frac{1}{R_c}\right)^2+\left(\omega_c C_c - \frac{1}{\omega_c L}\right)^2}} \qquad (14)$$

where R_c and C_c are functions of V_m and the biasing voltage V_0. Since R_c and C_c, the effective resistance and capacitance of the ferroelectric condenser are nonlinear, the voltage would contain harmonics of the carrier if the current i_0 were sinusoidal. However, we suppose that there is connected across the crystal a two-terminal network having a very high impedance to the carrier and a very low impedance to the harmonics (a simple anti-resonant circuit will approximate this); then the analysis can be carried out in a simple manner. Since

$$\frac{dV_m}{dV_0} = i_{0_m}\left[\frac{\partial |Z|}{\partial V_m}\frac{dV_m}{dV_0}+\frac{\partial |Z|}{\partial V_0}\right], \qquad (15)$$

we have

$$\frac{dV_m}{dV_0} = \frac{i_{0_m}\dfrac{\partial |Z|}{\partial V_0}}{1-\dfrac{V_m}{|Z|}\dfrac{\partial |Z|}{\partial V_m}}.$$

Using the definitions

$$\mu_0 \equiv \frac{V_m}{C_c}\frac{\partial C_c}{\partial V_0}; \qquad \mu_m \equiv \frac{V_m}{C_c}\frac{\partial C_c}{\partial V_m};$$

$$\beta_0 = V_m R_c \frac{\partial\left(\frac{1}{R_c}\right)}{\partial V_0}; \qquad \beta_m = V_m R_c \frac{\partial\left(\frac{1}{R_c}\right)}{\partial V_m}$$

$$\omega_c = 2\pi f_c; \qquad D \equiv 1+\frac{R_c}{R_2}; \qquad Q = \omega_c R_c C_c;$$

$$X = 1-\frac{1}{\omega_c^2 L C_c} = 1-\frac{\omega_t^2}{\omega_c^2}$$

we find, on performing the differentiations indicated in (15) on the form of $|Z|$ indicated in (14), that

$$G \equiv \frac{dV_m}{dV_0} = -\frac{\beta_0 D/Q^2+\mu_0 X}{D(D+\beta_m)/Q^2+\mu_m X+X^2}. \qquad (16)$$

A signal, low in frequency compared to the carrier, is equivalent to a varying bias voltage. Due to this signal the carrier voltage is amplitude modulated, and if a perfect detector is used which follows the envelope, the power output is given by

$$P_{\text{out}} = \frac{V_0^2 G^2}{R_1} \qquad (17)$$

where V_1 is the rms signal voltage applied to the crystal. The input circuit is preferably designed as a filter, as shown by Fig. 31(b), where C_0 is the capacitance of the crystal at the signal frequency.

The power input is then

$$P_{\text{in}} = \omega_S^2 C_0^2 V_0^2 R_1 \qquad (18)$$

and the power gain is

$$\frac{P_{\text{out}}}{P_{\text{in}}} = \frac{G^2}{R_1 R_2 \omega_S^2 C_0^2}. \qquad (19)$$

Using the filter design formula $R_1 = 1/\omega_0 C_0$, where $\omega_0 = 2\pi f_0$, where f_0 is the cut-off frequency of the filter, and the relation $R_c = Q/\omega_0 C_0$, we get

$$\frac{P_{\text{out}}}{P_{\text{in}}} = \frac{C_c}{C_0}\frac{R_c}{R_2}\frac{G^2}{Q}\frac{f_0 f_c}{f_S^2}. \qquad (20)$$

Factoring out the constant and frequency variable parts

of (16), we have

$$G = \frac{-\beta_0}{D+\beta_m}\left[\frac{1+\frac{\mu_0 Q^2}{\beta_0 D}X}{1+\frac{\mu_m Q^2}{D(D+\beta_m)}X+\frac{Q^2}{D(D+\beta_m)}X^2}\right]$$

$$= -K\left(\frac{1+AX}{1-BX+CX^2}\right). \tag{21}$$

From this equation we see that if we operate the amplifier at anti-resonance, i.e., $X=0$, all the modulation is caused by resistance modulation. Since μ_m, the change in capacitance with bias is negative, the term multiplying X in the denominator is written in this form on the right-hand side. With this substitution, the gain can be written as

$$\frac{P_{\text{out}}}{P_{\text{in}}} = \frac{C_c}{C_0}\frac{R_c}{R_2}\frac{K^2}{Q}\frac{f_0 f_c}{f_S^2}(\text{TF}) \tag{22}$$

where the tuning factor TF is the square of the factor of (21) in brackets.

To calculate the factors appearing in the above equations, we make use of the idealized hysteresis loops of Fig. 31(c). These are reasonably close to the values shown by Figs. 28 and 29. The effect of a bias is assumed to be to cut down the size of the loop as shown by the dashed line. Let V be the instantaneous voltage and q the charge. Then,

$$V = V_m \sin\theta + V_0 \quad \text{where} \quad \theta = \omega t. \tag{23}$$

For various parts of the loop the charge q can be written

$$\begin{aligned}
q &= (V+V_c)C_0';\ V_0-V_m<V<V_s-2V_c,\\
&= (V_s-V_c)C_0';\ V_s-2V_c<V<V_0+V_m,\\
&= (V_s-V_c)C_0';\ V_s<V<V_m+V_0,\\
&= (V-V_c)C_0';\ V_0-V_m+2V_c<V<V_s,\\
&= (V_0+V_c-V_m)C_0';\ V_0-V_m<V<V_0-V_m+2V_c
\end{aligned} \tag{24}$$

provided $V_0 > V_m - V_s$. As seen from Fig. 31(c), V_c is the coercive voltage, V_s is the width of the hysteresis loop at the bottom, and V_m is the complete swing of the carrier voltage. C_0' is the capacitance corresponding to the sloping parts of the loop and it is assumed that no additional charge is generated when the voltage is higher than V_s.

We now assume a Fourier series for q

$$q = q_0 + q_1 \sin\theta + p_1 \cos\theta + \text{harmonics}. \tag{25}$$

On account of the low impedance to the harmonics, these can be neglected. Integrating

$$q_0 = \frac{1}{2\pi}\int_{-\pi/2}^{3\pi/2} q\,d\theta; \qquad q_1 = \frac{1}{\pi}\int_{-\pi/2}^{3\pi/2} q\sin\theta\,d\theta;$$

$$p_1 = \frac{1}{\pi}\int_{-\pi/2}^{3\pi/2} q\cos\theta\,d\theta. \tag{26}$$

Note that since $V = V_m \sin\theta + V_0$; $dV = V_m\cos\theta d\theta$. Hence

$$p_1 = \frac{1}{\pi V_m}\int q\,dV = \frac{1}{\pi V_m}$$

times area of loop, so that

$$p_1 = \frac{2}{\pi}\frac{V_c}{V_m}(V_m + V_s - V_0 - 2V_c)C_0'. \tag{27}$$

Evaluation of q_1 gives the result

$$q_1 = \frac{V_m C_0'}{4\pi}[\pi + 2(\theta_1 - \theta_2 + \theta_3) + \sin 2\theta_1 - \sin 2\theta_2 + \sin 2\theta_3] \tag{28}$$

where

$$\sin\theta_1 = \frac{V_s}{V_m} - \frac{V_0}{V_m} - \frac{2V_c}{V_m}; \qquad -\frac{\pi}{2} < \theta_1 < \frac{\pi}{2}$$

$$\sin\theta_2 = \frac{V_s}{V_m} - \frac{V_0}{V_m}; \qquad +\frac{\pi}{2} < \theta_2 < \frac{3\pi}{2}$$

$$\sin\theta_3 = -1 + \frac{2V_c}{V_m}; \qquad \frac{\pi}{2} < \theta_3 < \frac{3\pi}{2}.$$

The values of $\theta_{1,2,3}$ are determined from the equations of Fig. 31(c) which are respectively

$$(e) = (V_s - 2V_c); \quad (c) = V_s; \quad (b) = 2V_c - V_m - V_0.$$

We also find by direct integration

$$q_0 = \frac{C_0'}{2\pi}\left[2\pi(V_c+V_0) - \frac{3\pi}{2}V_m + \theta_1(2V_c + V_0 - V_s) + \theta_2(V_s - V_0) + \theta_3(V_m - 2V_c) - V_m(\cos\theta_1 - \cos\theta_2 + \cos\theta_3)\right]. \tag{29}$$

Since by definition

$$C_c = \frac{q_1}{V_m} \quad \text{and} \quad \frac{1}{R_c} = \frac{\omega_c p_1}{V_m} \tag{30}$$

we find

$$C_c = \frac{C_0'}{4\pi}[\pi + 2(\theta_1 - \theta_2 + \theta_3) + \sin 2\theta_1 - \sin 2\theta_2 + \sin 2\theta_3] \tag{31}$$

$$\frac{1}{R_c} = \frac{\omega C_0'}{\pi}(1+\sin\theta_1)(1+\sin\theta_3)$$

and hence

$$Q \equiv \omega_c R_c C_c = \frac{\pi + 2(\theta_1 - \theta_2 + \theta_3) + \sin 2\theta_1 - \sin 2\theta_2 + \sin 2\theta_3}{4(1+\sin\theta_1)(1+\sin\theta_3)}. \tag{32}$$

By carrying out the differentiations indicated in their definitions, we get

$$\mu_m = -2\left[\frac{2\cos\theta_3+\sin 2\theta_1-\sin 2\theta_2+\sin 2\theta_3}{\pi+2(\theta_1-\theta_2+\theta_3)+\sin 2\theta_1-\sin 2\theta_2+\sin 2\theta_3}\right]$$

$$\mu_0 = \frac{4(\cos\theta_2-\cos\theta_1)}{\pi+2(\theta_1-\theta_2+\theta_3)+\sin 2\theta_1-\sin 2\theta_2+\sin 2\theta_3} \qquad (33)$$

$$\beta_m = -\frac{1+2\sin\theta_1}{1+\sin\theta_1}$$

$$\beta_0 = -\frac{1}{1+\sin\theta_1}.$$

The crystal capacitance at the signal frequency is defined by

$$C_0 \equiv \frac{dq_0}{dV_0} = \frac{\partial q_0}{\partial V_0} + \frac{\partial q_0}{\partial V_m}\frac{dV_m}{dV_0}. \qquad (34)$$

By differentiation, we find

$$\frac{\partial q_0}{\partial V_0} = \frac{C_0'}{2\pi}(2\pi+\theta_1-\theta_2);$$

$$\frac{\partial q_0}{\partial V_m} = \frac{C_0'}{2\pi}\left(\theta_3 - \frac{3\pi}{2} - \cos\theta_1 + \cos\theta_2 - \cos\theta_3\right).$$

Hence,

$$C_0 = \frac{C_0'}{2\pi}\left(2\pi+\theta_1-\theta_2 + G\left[\theta_3 - \frac{3\pi}{2} - \cos\theta_1 + \cos\theta_2 - \cos\theta_3\right]\right). \qquad (35)$$

This simple theory indicates that the resistance at the signal frequency associated with C_0 is small enough to neglect. Some experimental verification of this conclusion is shown by the curves of Fig. 30. The rise at the top of the modulation curve is due to the bias voltage plus signal voltage plus carrier voltage adding together in phase so that the total voltage has a value according to point *d* of Fig. 31. In this figure the line *e–d* is taken horizontal, but the actual hysteresis loop of Fig. 5 shows that the polarization rises with voltage. Thus the time corresponding to the maximum of the top curve is the time when the phases of the signal and carrier voltage are such that they add. At this same time the current should be a minimum if the resistance is small and, as seen from Fig. 30, these times coincide. If there were an appreciable resistance, the phase of the modulated current would differ from the phase of the voltage and a phase displacement would occur between the minimum current and the maximum voltage addition. Since this does not occur, we can assume the resistance effect to the signal is small.

Some general conclusions can be reached by carrying out calculations on these formulas for various parameters of the hysteresis loop. Some of these are:

1. The ideal characteristic is one having nonlinearity without hysteresis, i.e., hysteresis cuts down the gain and introduces heating.
2. The smaller the ratio of V_c/V_s the better the operation. For good crystals this ratio is under a half and the ratio of the gain obtained to the ideal gain, i.e., $K^2=4$ is about 0.4 to 0.5.
3. If V_c/V_s is one-half or less, the denominator of the tuning factor can be made small and some advantage can be obtained from this factor.

Design Considerations of Junction Transistors at Higher Frequencies

Based Upon an Accurate Equivalent Circuit*

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Summary—**The analytical solution of the diffusion equation has been used to construct an accurate equivalent circuit of the junction transistor. All network components are expressed in terms of the physical parameters of the transistor. This network has been used to calculate Mason's *U*-function, which in turn gives the highest frequency at which a transistor can give a power gain. From the derived formula, it is possible to see how the physical parameters interact to limit the frequency response. The design requirements for a given maximum frequency can be obtained from the formulas; a few examples are discussed. The accurate equivalent circuit will be helpful in designing transistors for more specialized applications.**

IN THE PAST, much of the work on junction transistors has been concerned with operation in the audio range. However, during the last year an increasing effort has been made to push the applicability of transistors to higher frequencies.[1–3] This has been done by essentially decreasing the base width, the col-

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[1] W. E. Bradley, "Principles of the surface-barrier transistor," Proc. I.R.E., vol. 41, pp. 1702–1706; December, 1953.

[2] J. M. Early, "P-N-I and N-P-I-N junction transistor triodes," *Bell Sys. Tech. Jour.*, vol. 33, pp. 517–533; May, 1954.

[3] C. W. Mueller and J. I. Pankove, "A *p-n-p* triode alloy-junction transistor for radio-frequency amplification," Proc. I.R.E., vol. 42, pp. 386–391; February, 1954.

lector capacitance, and the ohmic base resistance. It is difficult to decrease these quantities simultaneously. In the conventional transistor design, if one of these quantities is made very small, another becomes very large. Compromises become necessary; thus the relative importance of these quantities must be known.

The design problem of a transistor may be divided into three parts. First, the physical processes that determine the properties of the transistor must be understood, so that analytical or numerical results may be obtained for the electrical parameters. Secondly, these results should be re-expressed in as simple an equivalent circuit as possible, so that network calculations can be carried out. Every component of the equivalent network should be expressed as a function of the resistivities, geometry, and other physical parameters. Thirdly, this electrical circuit should be optimized for a specific application by the proper choice of these parameters. In this paper, the limitation of the frequency response by the physical parameters is investigated. The highest frequency at which a transistor can give a power gain is obtained. This calculation can be used to determine, among other things, how large the collector capacity and the base resistance can be made, in order to make full use of a small obtainable base width.

This work can easily be extended to transistors that deviate in some respect from conventional units. It is well known that the emitter capacity in conventional transistors is negligible; thus, it has been omitted in this derivation. However, the large emitter capacity of Early's *n-p-i-n* and *p-n-i-p* transistors[2] can be included, and their highest operating frequency can be calculated. The equivalent circuit given should also be helpful in answering many other questions concerning the operation at high frequencies.

Work on the first part of the design problem overlaps that of Early.[4] The presentation for the grounded base connection at which finally arrived is given in Fig. 1.

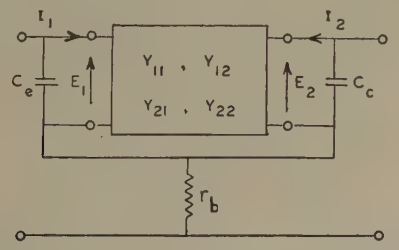

Fig. 1—Schematic representation of junction transistor in grounded base connection.

The following derivation for the equivalent circuit is based on the analytic expressions for the short-circuit admittances of a one-dimensional junction transistor. The point of departure in these studies is the schematic representation in Fig. 1. Units having a low zero frequency α and, therefore, a high dc base current will not be considered here. The voltage generator in series with r_b is neglected and the base resistance is not split into two components. C_c is the "collector capacitance," C_e is the "emitter capacitance," and r_b is the "base spreading" resistance.

The short-circuit driving-point and transfer admittances of the box are defined by the relations[5]

$$Y_{11} = gs_n + \frac{as_p}{\tanh s_p}, \qquad Y_{12} = -\frac{bs_p}{\sinh s_p},$$
$$Y_{21} = -\frac{as_p}{\sinh s_p}, \qquad Y_{22} = \frac{bs_p}{\tanh s_p}, \tag{1}$$

where

$$a = \frac{q}{kT} I_{pe}, \tag{2}$$

$$b = I_{pe} \frac{1}{W_o} \left| \frac{\partial W_o}{\partial V_c} \right|, \tag{3}$$

$$g = a \frac{n_n}{p_p} \left(\frac{D_n}{D_p} \times \frac{\tau_p}{\tau_n} \right)^{1/2}, \tag{4}$$

and

$$s_p = \frac{W_o}{L_p} (1 + j\omega\tau_p)^{1/2}; \qquad s_n = \frac{W_o}{L_p} (1 + j\omega\tau_n)^{1/2}. \tag{5}$$

In (1) through (5) the following symbols are used:

q = electronic charge,
k = Boltzmann's constant,
T = absolute temperature in degrees Kelvin,
I_{pe} = dc hole current flowing over emitter junction,
W_o = base width at dc operating bias,
V_c = dc collector bias,
n_n = equilibrium concentration of electrons in base region,
p_p = equilibrium concentration of holes in emitter region,
D_n, D_p = diffusion constants of electrons and holes, respectively,
τ_p = lifetime of a hole in the base region,
τ_n = lifetime of an electron in the emitter region.

Numerical values of the physical constants considered here are:

$$\left.\begin{aligned} q/kT &= 40 \text{ volt}^{-1}, \\ I_{pe} &= 10^{-3} \text{ amp}, \\ n_n/p_p &= 10^{-1} \text{ to } 10^{-3}, \\ W_o &\approx 10^{-3} \text{ cm}, \\ D_n &= 90 \text{ cm}^2/\text{sec}, \quad D_p = 43 \text{ cm}^2/\text{sec}, \\ \tau_n &= \tau_p = 10^{-4} \text{ sec}, \\ L_p &= (D_p \times \tau_p)^{1/2}, \\ \left| \frac{\partial W_o}{\partial V_c} \right| &\approx 2 \times 10^{-5}, \\ r_b &\approx 100 \text{ ohms}, \\ C_c &\approx 10 \text{ to } 100 \ \mu\mu\text{f}. \end{aligned}\right\} \tag{6}$$

Using these values, the constants a, b, and g become:

[4] J. M. Early, "Design theory of junction transistors," *Bell Sys. Tech. Jour.*, vol. 32, pp. 1271–1312; November, 1953.

[5] These equations are essentially those of J. M. Early; see ref. 4.

$$\left.\begin{aligned} a &= 4 \times 10^{-2}, \\ b &= 2 \times 10^{-5}, \\ g &= 4\sqrt{2}\ (10^{-3} \text{ to } 10^{-5}). \end{aligned}\right\} \quad (7)$$

The quantity s_p becomes (in the numerical example $s_p = s_n$, because $\tau_p = \tau_n$):

$$s_p = \frac{W_o}{(D_p\tau_p)^{1/2}}(1 + j\omega\tau_p)^{1/2} \approx \sqrt{2}\ 10^{-2}(1 + j\omega\tau_p)^{1/2}, \quad (8)$$

and for sufficiently high frequencies where $\omega\tau_p = 10^{-4}$ $\omega \gg 1$,

$$\begin{aligned} s_p &\approx \frac{W_o}{\sqrt{D_p}}(j\omega)^{1/2} \approx \sqrt{2}\ 10^{-4}(j\omega)^{1/2} \\ &= (1 + j)10^{-4}\sqrt{\omega}. \end{aligned} \quad (9)$$

In order to arrive at a useful network, representation for the box in Fig. 1, (1) may be written as

$$\begin{aligned} Y_{11} &= y + ny_{11}, & Y_{12} &= y_{12}, \\ Y_{21} &= ny_{21}, & Y_{22} &= y_{22}, \end{aligned} \quad (10)$$

with

$$y_{11} = y_{22} = \frac{bs_p}{\tanh s_p}, \quad (11)$$

$$y_{12} = y_{21} = -\frac{bs_p}{\sinh s_p}, \quad (12)$$

$$y = gs_n, \quad (13)$$

$$n = a/b \approx 2000. \quad (14)$$

A two-terminal-pair network is defined with the properties illustrated in Fig. 2. Like an ideal transformer with a ratio n, the input voltage is changed by this factor.

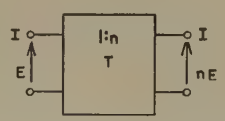

Fig. 2—Two-terminal-pair network representing the active element in the transistor.

Unlike an ideal transformer, however, the current remains unchanged so that the output power is n times the input power. Also, an impedance Z across the output looks like an impedance Z/n at the input, and Z placed across the input looks like nZ when viewed from the output terminals. Circuit-wise this network is as simple to work with as an ideal transformer. From a physical standpoint, it is an active element and represents completely both the active and the non-reciprocal properties of the transistor.[6] The resulting circuit, except for the component shown in Fig. 2, is an ordinary linear, passive, bilateral network.

[6] The same separation into active and passive parts was suggested by J. Zawels at the Transistor Circuit Conference at Philadelphia in February, 1954.

Thus (10), (11), and (12) show that the box in Fig. 1 has the schematic representation given in Fig. 3, in which the linear, passive, bilateral, and symmetrical two-terminal-pair is characterized by the short-circuit admittances of (11) and (12). This network may tentatively be represented physically as a symmetrical lattice with the component admittances

$$y_a = y_{11} - y_{12} = bs_p\frac{\cosh s_p + 1}{\sinh s_p} = bs_p \coth s_p/2, \quad (15)$$

$$y_b = y_{11} + y_{12} = bs_p\frac{\cosh s_p - 1}{\sinh s_p} = bs_p \tanh s_p/2, \quad (16)$$

or with the impedances

$$z_a = \frac{\tanh s_p/2}{bs_p}, \qquad z_b = \frac{\coth s_p/2}{bs_p}. \quad (17)$$

Appropriate network schematics must still be found to represent the box y in Fig. 3 as well as the symmetrical lattice with the impedances of (17).

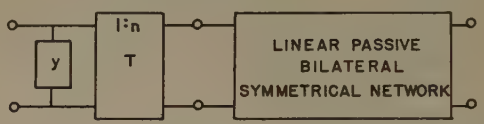

Fig. 3—Schematic network representation of the box of Fig. 1.

For the box y and (8) and (13), it is essential to find an appropriate rational approximation to the irrational expression $(1+j\omega\tau_n)^{1/2}$. Consider the squared magnitude of the function $(1+j\bar{\omega})^{1/2}$, which is $(1+\bar{\omega}^2)^{1/2}$, and plot the logarithm of this squared magnitude versus the logarithm of $\bar{\omega}$ (Fig. 4). This is essentially a straight line ris-

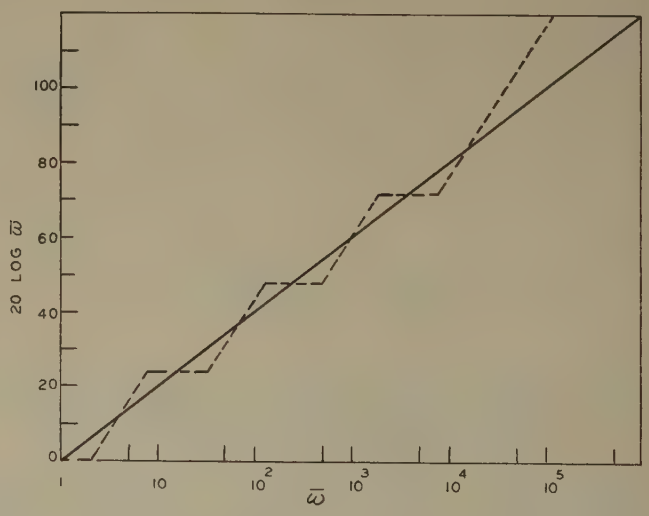

Fig. 4—The squared magnitude of $(1+j\bar{\omega})^{1/2}$ in decibels versus $\bar{\omega}$.

ing with a slope of 20 db per decade, for $20 \log (1+\bar{\omega}^2)^{1/2} \approx 20 \log \bar{\omega}$. A rational approximation to the squared magnitude of $(1+j\bar{\omega})^{1/2}$ has the form

$$\frac{(1 + \bar{\omega}^2/\bar{\omega}_1^2)(1 + \bar{\omega}^2/\bar{\omega}_3^2)\cdots}{(1 + \bar{\omega}^2/\bar{\omega}_2^2)(1 + \bar{\omega}^2/\bar{\omega}_4^2)\cdots}. \quad (18)$$

Taking 20 log of this expression, the asymptotic slope per factor is 40 db per decade. Hence an approximation may be obtained by alternately rising at the rate of 40 db per decade and zero db per decade, as shown by the dotted line in Fig. 4. Since the actual curve will be smooth and deviate ± 3 db from the dotted line, an approximation that has a maximum deviation of 6 db will yield a Tchebycheff-like approximation with a 3 db ripple.

The break points in this approximation occur at log $\bar{\omega} = 0.3, 0.9, 1.5, 2.1, 2.7, 3.3, 3.9$, yielding the critical frequencies: $\bar{\omega}_k = 2, 8, 31.6, 126, 500, 2000, 8000$; hence the rational approximation

$$(1+\bar{s})^{1/2} \approx \frac{(\bar{s}+2)(\bar{s}+31.6)(\bar{s}+500)(\bar{s}+8000)}{125(\bar{s}+8)(\bar{s}+126)(\bar{s}+2000)} = y^*. \quad (19)$$

The frequency variable $\bar{s}$ in this expression is to be identified with $j\omega\tau_n$ and, since $\tau_n = 10^{-4}$, the highest critical frequency $\omega\tau_n$ in the above rational expression is 8×10^7, corresponding to $\omega = 13$ mc. For a lower lifetime material, one would need fewer terms in the expression; for a higher lifetime τ_n, more terms are required.

A network for the admittance function (19) is readily found by expanding $y^*/\bar{s}$ in partial fractions and multiplying by $\bar{s}$ to give

$$y^* = 1 + \frac{2.36\bar{s}}{\bar{s}+8} + \frac{9.85\bar{s}}{\bar{s}+126} + \frac{37.8\bar{s}}{\bar{s}+2000} + \frac{\bar{s}}{125}. \quad (20)$$

The corresponding network is shown in Fig. 5.[7]

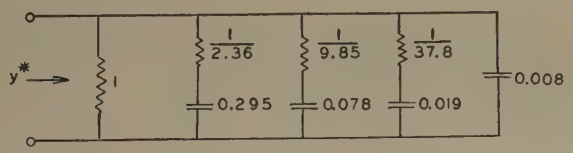

Fig. 5—The normalized equivalent network of the y-box.

Expanding the frequency scale by the factor $1/\tau_n = 10^4$, the capacitance values in this circuit become multiplied by $\tau_n = 10^{-4}$. The admittance

$$y = \frac{q}{kT} I_{p_0} \frac{n_n}{p_p} \left(\frac{D_n}{D_p} \times \frac{\tau_p}{\tau_n}\right)^{1/2} \frac{W_o}{L_p} y^*$$
$$= (10^{-4} \text{ to } 10^{-6}) y^*,$$

and, choosing the intermediate value 10^{-5} for this amplitude factor, the network for y is obtained by multiplying the resistances in Fig. 5 by 10^5 and the capacitances by the additional factor 10^{-5}. The circuit of Fig. 6 is thus obtained for y. The dotted lines indicate points where this network may be cut, and still be an approximate realization for y with the same tolerance when the highest frequency to be considered is less than 50 mc.

Next, the admittance y must be associated with Fig. 3. To do this, y must be divided by n and placed on the

[7] Units of resistance, capacitance, and inductance in the following circuit diagrams are in ohms, farads, and henries, respectively.

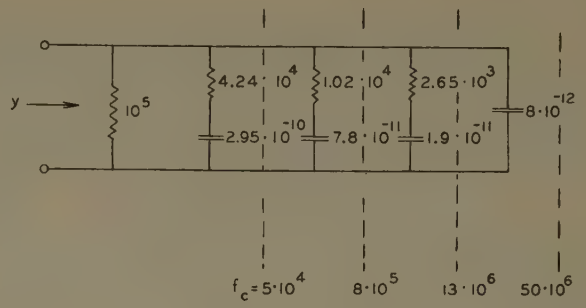

Fig. 6—The equivalent network of the y-box for the given numerical example.

right-hand side of the network T. Since it will be found expedient in the synthesis of the symmetrical network of Fig. 3 to consider the impedances (17) on a level of $2b$ ohms, the admittance y of Fig. 6 should be divided by $2bn = 2a = 8/100$. In the numerical example it then takes the form shown in Fig. 7.

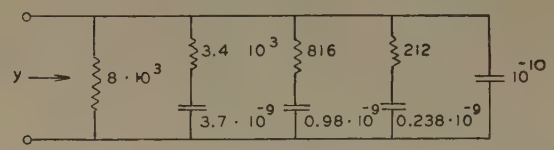

Fig. 7—The circuit of Fig. 6 placed on the right-hand side of the active T-network (on an impedance level of $2b$ ohms).

If the symmetrical network of Fig. 3 is considered in the form of a lattice on a $2b$-ohm impedance level, and the collector capacitance is associated with it, the resultant circuit may be converted into an equivalent T-network as shown in Fig. 8.

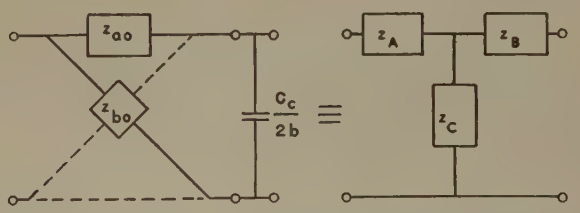

Fig. 8—Conversion of the lattice with associated collector capacity C_c to an equivalent T-network.

The impedance of the collector capacitance is

$$Z_2 = \frac{2b}{C_c s}, \quad \text{where} \quad s = j\omega, \quad (21)$$

and the lattice impedances according to (17) are

$$z_{ao} = 2bz_a = \frac{\tanh s_p/2}{s_p/2}; \quad z_{bo} = 2bz_b = \frac{\coth s_p/2}{s_p/2}. \quad (22)$$

The impedances in the equivalent T-circuit are given by the expressions

$$z_A = \frac{z_{ao}z_{bo} + z_{ao}Z_2}{\frac{1}{2}(z_{bo} + z_{ao}) + Z_2}, \tag{23}$$

$$z_B = \frac{z_{ao}Z_2}{\frac{1}{2}(z_{bo} + z_{ao}) + Z_2}, \tag{24}$$

$$z_C = \frac{\frac{1}{2}(z_{bo} - z_{ao})Z_2}{\frac{1}{2}(z_{bo} + z_{ao}) + Z_2}. \tag{25}$$

Using (22) and (23), and introducing the factor

$$K = \frac{D_pC_c}{bW_o^2} = \frac{D_pC_c}{W_oI_{pe}\left|\frac{\partial W_o}{\partial V_c}\right|} \approx 21.5 \times 10^{11}\, C_c, \tag{26}$$

these impedances may be written in the form

$$z_A = \frac{2K + \frac{\tanh s_p/2}{s_p/2}}{1 + Ks_p \coth s_p}, \tag{27}$$

$$z_B = \frac{\frac{\tanh s_p/2}{s_p/2}}{1 + Ks_p \coth s_p}, \tag{28}$$

$$z_C = \frac{\frac{2}{s_p \sinh s_p}}{1 + Ks_p \coth s_p}. \tag{29}$$

The poles of these functions lie where

$$Ks_p + \tanh s_p = 0. \tag{30}$$

Except for z_C, which also has a pole at $s_p=0$, the three impedances have identical poles given by

$$\tanh s_p = -Ks_p. \tag{31}$$

Letting

$$s_p = ju, \tag{32}$$

the transcendental equation (31) is transformed into the more easily interpreted form

$$\tan u = -Ku. \tag{33}$$

In considering the roots of this equation, it is significant to note from (26) for K, that, if the collector capacitance C_c is assumed to lie within the range 10^{-11} to 10^{-10} farad, then K has values from 21.5 to 215. Even if the smallest of these values is taken, the roots of (33) are approximately odd multiples of $\pi/2$. For example, if the smallest root (for which the deviation is greatest) assumes the value $(\pi/2)+\delta$, then (33) yields

$$\tan\left(\frac{\pi}{2} + \delta\right) \approx -\frac{1}{\delta} = -K\left(\frac{\pi}{2} + \delta\right);$$

$$\delta \approx \frac{2}{K\pi}. \tag{34}$$

With $K=21.5$, this gives $\delta=0.030$ or a 3 per cent deviation.

The pole positions for the impedances (27), (28), and (29) are thus almost independent of the value of C_c within its normal range, and are given by

$$u_\nu = \frac{\nu\pi}{2} \quad \text{for} \quad \nu = 1, 3, 5, \cdots. \tag{35}$$

Let a new frequency variable p be defined as

$$p = \frac{4}{\pi^2}\frac{W_o^2}{D_p}s = 0.97\frac{s}{\omega_\alpha} \approx \frac{s}{\omega_\alpha}, \tag{36}$$

where ω_α is the so-called cut-off frequency. (At this frequency, the ratio of the hole current over the collector junction to the hole current over the emitter junction has dropped to 0.707 of its low-frequency value. As can be shown, $\omega_\alpha=2.4\ D_p/W_o^2$.)

From (9), (32), and (36),

$$ju = s_p = \frac{W_o}{\sqrt{D_p}}\sqrt{s} = \frac{\pi}{2}\sqrt{p}. \tag{37}$$

All the poles of the impedance functions are seen to lie on the negative real axis of the complex frequency plane at the points

$$p_\nu = -\nu^2 \quad \text{for} \quad \nu = 1, 3, 5, \cdots, \tag{38}$$

or at

$$p_\nu = -1, -9, -25, \cdots.$$

It will be found expedient to express the impedances z_A, z_B, z_C in terms of their partial fraction expansions. To this end, the residues of these impedances in their poles must be computed. In this regard,

$$\frac{d}{du}(1 + jKu \coth ju) \times \frac{du}{dp}\Big]_{u=\nu\pi/2}$$

$$= \frac{d}{du}(1 + Ku \cot u) \times \frac{du}{dp}\Big]_{u=\nu\pi/2}$$

$$\approx (-Ku) \times \left(\frac{-\pi^2}{8u}\right)\Big]_{u=\nu\pi/2} = \frac{K\pi^2}{8}. \tag{39}$$

Thus

$$\text{residues of } z_A = \frac{2K + \frac{\tan \nu\pi/4}{\nu\pi/4}}{K\pi^2/8}$$

$$= \frac{2K + 4(-1)^{(\nu-1)/2}/\nu\pi}{K\pi^2/8} \approx \frac{16}{\pi^2} \tag{40}$$

because $2K \gg |4/\nu\pi|$ for $\nu=1, 3, 5, \cdots$. Next

$$\text{residues of } z_B = \frac{\frac{\tan \nu\pi/4}{\nu\pi/4}}{K\pi^2/8} = \frac{32(-1)^{(\nu-1)/2}}{K\pi^3\nu}, \tag{41}$$

and

$$\text{residues of } z_C = \frac{-2}{(\nu\pi/2)\sin \nu\pi/2} \Big/ \frac{K\pi^2}{8} = \frac{-4(-1)^{(\nu-1)/2}}{\nu\pi} \Big/ \frac{K\pi^2}{8}$$

$$= \frac{32(-1)^{(\nu+1)/2}}{K\pi^3\nu}. \tag{42}$$

In addition for $s_p \to o$ or $p \to o$, the impedance z_C, by (29) and (37), becomes

$$z_C \to \frac{2/s_p^2}{1+K} = \frac{2}{1+K}\,\frac{4}{\pi^2}\,\frac{1}{p},$$

from which it is clear that z_C also has a pole at $p=o$ with the residue

$$\frac{8}{\pi^2(1+K)} \approx \frac{8}{K\pi^2}. \tag{43}$$

The impedances (27), (28), and (29) may thus be written

$$z_A = \frac{16}{\pi^2}\left\{\frac{1}{p+1}+\frac{1}{p+9}+\frac{1}{p+25}+\cdots\right\}; \tag{44}$$

$$z_B = \frac{32}{K\pi^3}\left\{\frac{1}{p+1}-\frac{1/3}{p+9}+\frac{1/5}{p+25}-\cdots\right\}; \tag{45}$$

$$z_C = \frac{32}{K\pi^3}\left\{\frac{\pi/4}{p}-\frac{1}{p+1}+\frac{1/3}{p+9}-\frac{1/5}{p+25}+\cdots\right\}. \tag{46}$$

It is useful to note that

$$z_A(o)=2;\ z_B(o)=\frac{1}{K};\ \left(z_C-\frac{8}{K\pi^2 p}\right)\Big]_{p=o} = -\frac{1}{K}. \tag{47}$$

If the infinite expansions (44), (45), and (46) are terminated at some finite point, the approximation will be improved if a constant term appropriate to the requisite behavior in the vicinity of zero frequency is added.

In this connection, according to (36), a frequency range from $\omega=o$ to $m\omega_\alpha$ corresponds to a p range from o to jm. For most purposes, therefore, only a few terms in the above partial fraction expansions will suffice to obtain an equivalent circuit representation with very good accuracy up to several times cut-off frequency.

For example, using three terms,

$$z_A = \frac{16}{\pi^2}\left\{\frac{1}{p+1}+\frac{1}{p+9}+\frac{1}{p+25}+0.083\right\}; \tag{48}$$

$$z_B = \frac{32}{K\pi^3}\left\{\frac{1}{p+1}-\frac{1/3}{p+9}+\frac{1/5}{p+25}-0.002\right\}; \tag{49}$$

$$z_C = \frac{32}{K\pi^3}\left\{\frac{\pi/4}{p}-\frac{1}{p+1}+\frac{1/3}{p+9}-\frac{1/5}{p+25}+0.002\right\}. \tag{50}$$

The equivalent circuit for the transistor (except for the box T and the admittance y of Fig. 3, and the base resistance $2br_b$) has, for three terms, the appearance shown in Fig. 9.

It is interesting to note that the sum of z_B and z_C reduces to the capacitive reactance $8/K\pi^2 p = 2b/C_c s$ of the collector junction alone. Comparison with the circuit in Fig. 7 for the admittance y shows that the latter is entirely negligible for most practical purposes. In cases where it is not negligible, it can easily be taken into account.

In order to return to an ordinary impedance level, the values of all resistances must be divided by $2b$ and the values of all capacities multiplied by $2b$. In addition, by returning to the ordinary frequency scale, the capacities must be divided by ω_α. Table I gives the factors by which the resistances and capacitances of Fig. 9 must be multiplied in the completely denormalized circuit.

TABLE I

Branch	z_A	z_B	z_C
Resistances	$\frac{8}{\pi^2}\,\frac{1}{b}$	$\frac{16}{\pi^3}\,\frac{1}{bK}$	$\frac{16}{\pi^3}\,\frac{1}{bK}$
Capacitances	$\frac{\pi^2}{8}\,\frac{b}{\omega_\alpha}$	$\frac{\pi^3}{16}\,\frac{Kb}{\omega_\alpha}=0.81C_c$	$\frac{\pi^3}{16}\,\frac{Kb}{\omega_\alpha}=0.81C_c$

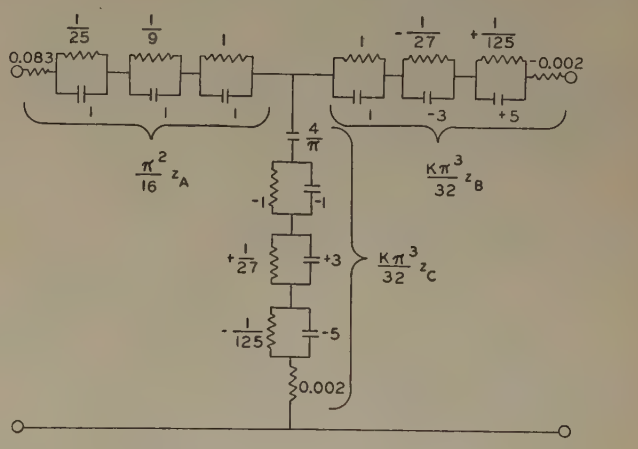

Fig. 9—The explicit form of the T-network of Fig. 8 (three terms in the approximation).

This equivalent circuit will be used to compute the unilateral gain[8] of the transistor as a function of frequency and of its physical parameters. In terms of the over-all (open-circuit driving-point and transfer) impedances Z_{11}, Z_{22}, Z_{12}, and Z_{21}, and their real parts R_{11}, R_{22}, R_{12}, and R_{21}, this quantity is defined by

$$U = \frac{|Z_{12}-Z_{21}|^2}{4(R_{11}R_{22}-R_{12}R_{21})}. \tag{51}$$

Since $Z_{12}=z_C/n+2br_b$ and $Z_{21}=z_C+2br_b$,

$$|Z_{12}-Z_{21}|^2 = \left(\frac{n-1}{n}\right)^2 |z_C|^2 \approx |z_C|^2. \tag{52}$$

[8] S. J. Mason, "Power Gain in Feedback Amplifiers," Technical Report No. 257, R.L.E., Mass. Inst. Tech.; August, 1953.

To compute the pertinent real parts, first,

$$z_A = r_A + jx_A, \quad z_B = r_B + jx_B, \quad z_C = r_C + jx_C, \tag{53}$$

then

$$\begin{aligned} R_{11} &= \frac{r_A + r_C}{n} + 2br_b, & R_{12} &= \frac{r_C}{n} + 2br_b, \\ R_{22} &= r_B + r_C + 2br_b, & R_{21} &= r_C + 2br_b, \end{aligned} \tag{54}$$

hence

$$R_{11}R_{22} - R_{12}R_{21} = \frac{1}{n}(r_Ar_B + r_Ar_C + r_Br_C) + 2br_b\left(\frac{r_A}{n} + r_B\right). \tag{55}$$

Since $r_B+r_C=0$, it follows that $r_Ar_B+r_Ar_C+r_Br_C=r_Br_C=-r_B^2=-r_C^2$. But r_B^2 or r_C^2 is proportional to $1/K^2$, and terms in $1/K^2$ were neglected in determining the poles of the functions z_A, z_B, z_C. Consistency with this approximation requires that the term $(r_Ar_B+r_Ar_C+r_Br_C)$ be dropped therefore

$$U = \frac{|z_C|^2}{8br_b\left(\frac{r_A}{n} + r_B\right)} = \frac{r_C^2 + x_C^2}{8br_b\left(\frac{r_A}{n} + r_B\right)}. \tag{56}$$

With $p=jx$ we readily find from (49), (50), and (51),

$$r_A = \frac{16}{\pi^2}\left[\frac{1}{1+x^2} + \frac{1/9}{1+x^2/81} + \frac{1/25}{1+x^2/625} + 0.083\right], \tag{57}$$

$$r_B = -r_C = \frac{32}{K\pi^3}\left[\frac{1}{1+x^2} - \frac{1/27}{1+x^2/81} + \frac{1/125}{1+x^2/625} - 0.002\right], \tag{58}$$

$$x_C = \frac{-8}{K\pi^2 x}\left[1 - \frac{4}{\pi}\left(\frac{x^2}{1+x^2} - \frac{x^2/243}{1+x^2/81} + \frac{x^2/3125}{1+x^2/625}\right)\right]. \tag{59}$$

Introducing the functions

$$X_1(x) = \frac{-\pi^2}{8} Kx_Cx = 1 - \frac{4}{\pi}\left[\frac{x^2}{1+x^2} - \frac{x^2/243}{1+x^2/81} + \frac{x^2/3125}{1+x^2/625}\right], \tag{60}$$

$$R_1(x) = Kr_B = -Kr_C = \frac{32}{\pi^3}\left[\frac{1}{1+x^2} - \frac{1/27}{1+x^2/81} + \frac{1/125}{1+x^2/625} - 0.002\right], \tag{61}$$

$$R_2(x) = \frac{r_A}{2} = \frac{8}{\pi^2}\left[\frac{1}{1+x^2} + \frac{1/9}{1+x^2/81} + \frac{1/25}{1+x^2/625} + 0.083\right], \tag{62}$$

then

$$\left(\text{noting from (37) that } x = \frac{4}{\pi^2}\frac{W_o^2\omega}{D_p}\right)$$

U becomes

$$U = \frac{D_p}{2W_o^2r_bC_c\omega^2}F(x) = \frac{\omega_\alpha}{4.8r_bC_c\omega^2}F(x), \tag{63}$$

where

$$F(x) = \frac{X_1^2 + \frac{\pi^4}{64}x^2R_1^2}{R_1 + \frac{2K}{n}R_2}. \tag{64}$$

This factor equals $(1+2K/n)^{-1}\approx 1$ for $x=O$ (since $X_1(o)=R_1(o)=R_2(o)=1$) and depends upon physical parameters only through the quantity

$$\frac{K}{n} = \frac{Kb}{a} = \frac{D_pC_c}{W_o^2\frac{q}{kT}I_{pe}} = \frac{\omega_\alpha C_e}{2.4\frac{q}{kT}I_{pe}}, \tag{65}$$

since the functions X_1, R_1, R_2 are pure numerics.

Once the factor $F(x)$ is plotted versus x for several values of K/n, the unilateral gain of the transistor as a function of frequency, and for any values of its physical constants, is readily computed.

The functions X_1, R_1, and R_2 have been plotted in Figs. 10 to 12 as a function of $x\approx\omega/\omega_\alpha$ using one to three terms in the expansions (44) to (46). It can be seen that all three approximations agree well up to $x=2.5$ or $\omega\approx 2.5\omega_\alpha$. For most practical purposes the approximation using the first term only will therefore be adequate. The function $F(x)$ depends on the parameter K/n. $F(x)$ is plotted in Figs. 13 to 15 for the values $K/n=1.075\times10^{-2}$, 2.5×10^{-2}, and 10.75×10^{-2} using all three approximations. Again, the first approximation is adequate up to $x=2.5$ or $\omega\approx 2.5\omega_\alpha$. Furthermore, it is interesting to note that $F(x)$ is very close to unity for $K/n=1.075\times10^{-2}$ and 2.5×10^{-2}, up to $w=3\omega_\alpha$. For $K/n=10.75\times10^{-2}$, $F(x)$ is still reasonably constant over the same frequency range, but its value varies from 0.8 to 0.7. In the numerical example given, the value of $K/n=2.5\times10^{-2}$ corresponds to a collector capacity of about 23 $\mu\mu$f. Remembering that the base width in the example was 10^{-3} cm, it becomes obvious that for many practical cases the assumption $F(x)=1$ is a sufficient approximation.

As has been outlined,[8] the U-function yields the highest frequency at which a transistor in any connection, and in any external circuit can give a power gain. The

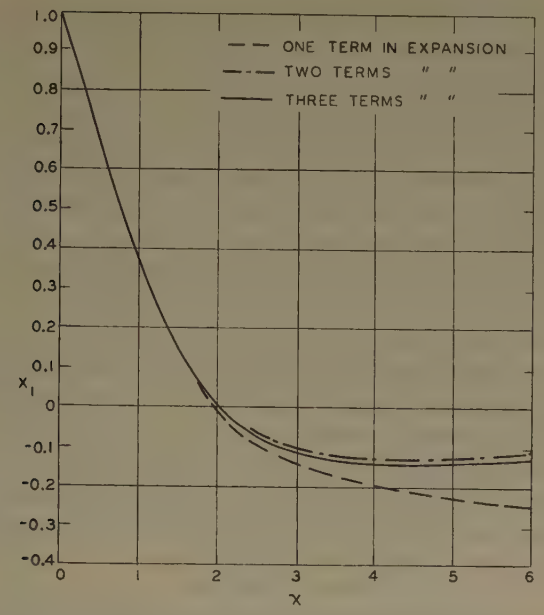

Fig. 10—The function X_1 versus $x \approx \omega/\omega_\alpha$.

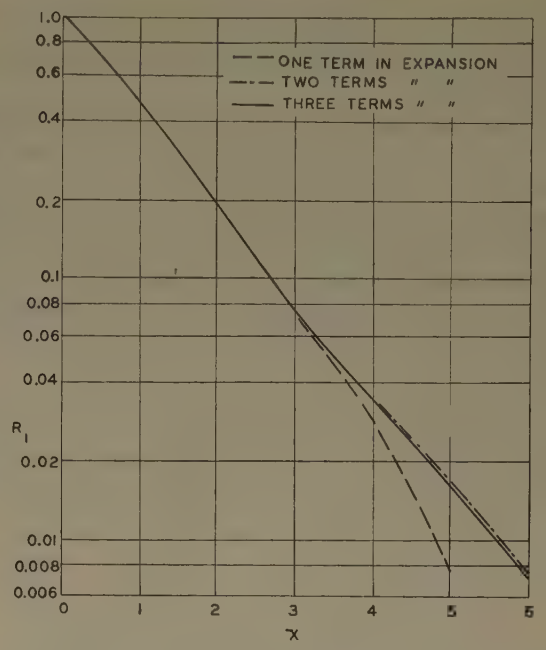

Fig. 11—The function R_1 versus $x \approx \omega/\omega_\alpha$.

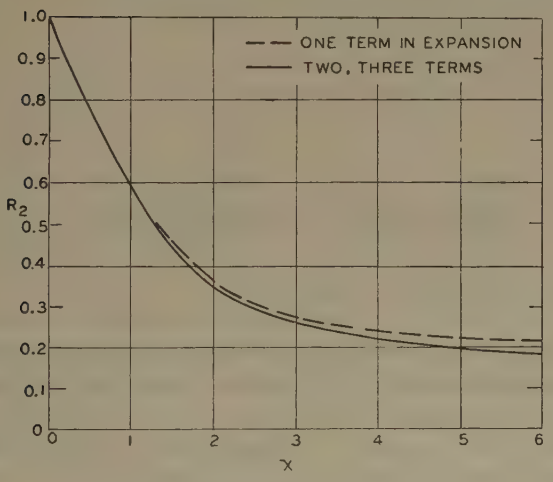

Fig. 12—The function R_2 versus $x \approx \omega/\omega_\alpha$.

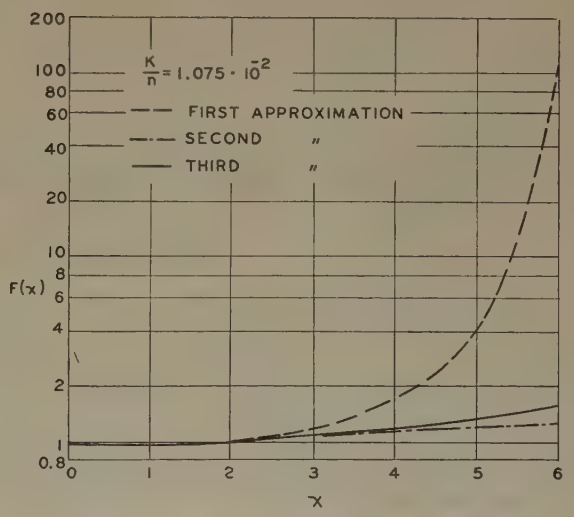

Fig. 13—The function $F(x)$ versus $x \approx \omega/\omega_\alpha$ for $K/n = 1.075 \times 10^{-2}$.

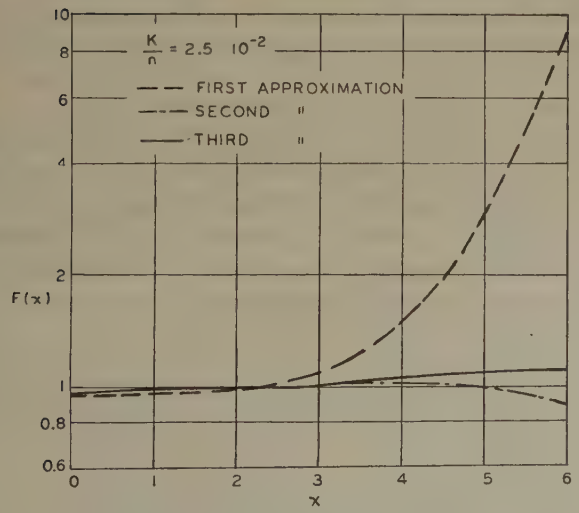

Fig. 14—The function $F(x)$ versus $x \approx \omega/\omega_\alpha$ for $K/n = 2.5 \times 10^{-2}$.

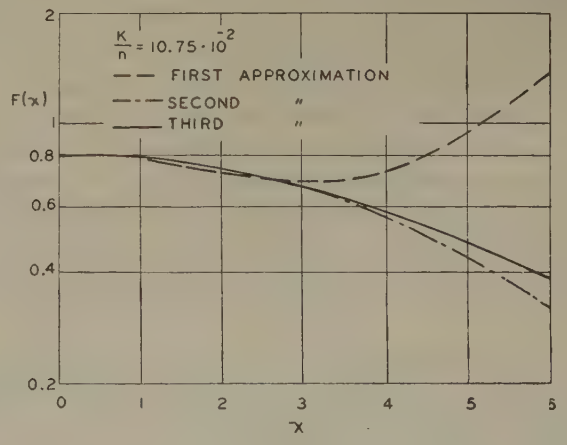

Fig. 15—The function $F(x)$ versus $x \approx \omega/\omega_\alpha$ for $K/n = 10.75 \times 10^{-2}$.

restrictions given there[8] do not apply here. The frequency at which U is equal to unity represents the limiting frequency.

Consider the following example. In a practical circuit, for a transistor to be used up to the cut-off frequency ω_α, the gain U must go through unity at a frequency higher than ω_α. Let this frequency be $\omega = 2\omega_\alpha$, and $F(x)$ assumed to be one. Table II gives values of the product r_bC_c for various frequencies f_α. C_c is determined by choosing a reasonable value of r_b in the conventional transistor designs.

TABLE II

f_α	r_bC_c (ohms f)	Assumed r_b (ohms)	Calculated C_c ($\mu\mu$f)
1 mc	8.3×10^{-9}	100	83
10 mc	8.3×10^{-10}	300	2.8
100 mc	8.3×10^{-11}	1000	0.083
1000 mc	8.3×10^{-12}	3000	0.0028

From Table II it becomes apparent that the conventional transistor will have its practical limit in the range of 100 mc. A fused collector with 0.083 $\mu\mu$f on 1 ohm cm n-type material at 6 v bias would have a diameter of 4.3×10^{-3} cm. The p-n-i-p and n-p-i-n transistor suggested by Early[2] might go to higher frequencies. These transistors were not discussed at the present time because the effect of the emitter capacity (which is not negligible in the above-mentioned transistor) was not included. Table II is correct only if throughout the entire range of values, the simplifying assumptions $F\alpha = 1$ and $K \gg 1$ are justified. It can be shown that $K \gg 1$ means small injection levels for alloy-junction transistors. For extremely small cross sections (which are required for very small capacities), the small injection condition is easily violated.

It is interesting to note that the quantity $|\partial W_o/\partial V_c|$ does not appear in the final expression for U and, therefore, has no influence on the highest operating frequency.

ACKNOWLEDGMENT

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Rippled Wall and Rippled Stream Amplifiers*

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Summary—**Two new single-velocity electron stream amplifiers, which use neither slow-wave guiding nor resonant circuitry, are described. The major part of the amplification is obtained in a drift-tube region where either the tube diameter or the stream diameter varies periodically. The period length is a quarter-plasma wavelength; the gain mechanism is one of reinforcing the simple harmonic motion of the electrons in the stream. A general theory is presented from which a gain formula is obtained for rippled wall, rippled stream, and velocity jump amplification. Design curves are shown. An experimental rippled wall amplifier operating in the range 1 to 4 kmc is discussed.**

INTRODUCTION

DURING THE past few years, there has been considerable discussion concerning the possibility of using a single stream of electrons without slow-wave guiding or resonant circuitry to obtain amplification. The possibility of obtaining growing waves on such a stream has been found to exist for nearly all simple nonpropagating circuits[1] but not for capacitive walls or short circuits, such as are provided by simple metal drift tubes. However, one can obtain amplification from a single stream in a drift tube by being just a bit careful about the relative disposition of the drift tube and stream. The essential condition is the periodic change, each quarter cycle of plasma oscillation of the stream, of this frequency or of the stream potential.[2]

One recalls that in a simple electron stream the individual charges perform simple harmonic motion about a uniformly moving mean position.[3,4] The realization of

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[1] C. K. Birdsall and J. R. Whinnery, "Waves in an electron stream with general admittance walls," *Jour. Appl. Phys.*, vol. 24, pp. 314–323; March, 1953.

[2] After our work was essentially completed, it was learned that similar theories were being worked on at several other laboratories, a situation not uncommon in this field. R. Kompfner, in England, had considered a sinusoidally rippled stream (1951, unpublished). R. C. Fletcher, Bell Telephone Labs., had made an analysis similar to that given here (1951, unpublished). H. Heffner, also at Bell, considered sinusoidally varying-stream parameters (1953, unpublished). S. Bloom and R. W. Peter, RCA Labs., presented "Transmission-line analog of a modulated electron beam," at the 1953 Conference on Electron Tube Research, subsequently published in the *RCA Rev.*, vol. 15, pp. 95–112; March, 1954. Peter, Bloom, and J. A. Ruetz published "Space-charge-wave amplification along an electron beam by periodic change of the beam impedance," same issue *RCA Rev.*, pp. 113–120. Bloom has made an analysis (unpublished) similar to that of Heffner. T. G. Mihran, GE Res. Labs., presented "Scalloped beam amplification," and J. A. Rich, same laboratory, presented "Space-charge waves in an electron beam of sinusoidally varying charge density," an analysis similar to that of Heffner and Bloom, at the 1954 Conference on Electron Tube Research. The only experimental work appears to be that of Mihran and the author.

[3] W. C. Hahn, "Small signal theory of velocity-modulated electron beams," *Gen. Elec. Rev.*, vol. 42, pp. 258–270; June, 1939.

[4] S. Ramo, "The electronic wave theory of velocity modulation tubes," PROC. I.R.E., vol. 27, pp. 757–763; December, 1939.

amplification can be made evident through analogy to a familiar simple harmonic oscillator, such as a pendulum. If the pendulum swings down in a strong (gravitational) field and up in a weak field, then the amplitude of swing must increase. Or, analogous to change in potential, if the pendulum's length (or mass) is increased on the swing down and decreased on the swing up, then amplitude must also increase; here the change may be essentially instantaneous, as contrasted to the field case. Thus, in the electron stream model, if the electrons could be made to see an alternating restoring force, or experience a potential jump, then their excursion from their mean position should increase, constituting amplification of an initial disturbance. Because the rate of increase is proportional to the excursion, the maximum excursion (at the end of each half-cycle) increases exponentially with distance. The field switching or potential jumping results simply from the electron's average velocity carrying it from one region to the next.

It appears that not-so-sudden changes in field or potential could also lead to amplification, but, one would guess, of a smaller amount than for sudden changes or jumps. Also, changes in field and in potential may be used in combination. Lastly, reversing the phase of the field or potential changes which produced amplification should lead to damping of the motion, or attenuation.

How may these field or potential changes be introduced into the electron stream model? The case of potential changes or velocity jumps has already been treated by Field, Tien, and Watkins.[5,6] The case of field changes is newer, requiring further comment. In this case, one varies the restoring force on the electrons in two relatively distinct ways: (1) by causing the stream to be close to a conducting wall, then farther from it, so that positive image charges in the wall alternately decrease then increase the restoring ("debunching") force; (2) by causing the stream to be large in diameter, then smaller. The stream debunches when it is small and of high current density; it bunches when it is large and of low current density.

A model for the first kind could be one in which the stream is maintained at constant diameter but moves through a drift tube of varying diameter; it is thus given the name *rippled wall amplifier*. A model for the second kind could be one in which the stream diameter is made to vary periodically (the stream envelope being stationary in space) with a drift tube of constant diameter not necessarily close to the stream; this kind is named the *rippled stream amplifier*. Combinations of the two kinds are possible; velocity jumps may be combined with either or both, and circuits other than drift tubes may be employed. A brief theory for the rippled wall amplifier was presented by J. R. Whinnery and the writer,[7] there called the *reduction factor amplifier*; here this theory will be elaborated, design curves presented, and experimental results given. The rippled stream amplifier will be treated briefly.

[5] L. M. Field, P. K. Tien, and D. A. Watkins, "Amplification by acceleration and deceleration of a single-velocity stream," PROC. I.R.E., vol. 39, p. 194; February, 1951.

[6] P. K. Tien, L. M. Field, "Space-charge waves in an accelerated electron stream for amplification of microwave signals," PROC. I.R.E., vol. 40, pp. 694–695; June, 1952.

SIMPLE THEORY

The model to be analyzed, shown in Fig. 1(a), consists of an electron gun, three rf sections, and a collector. The amplifying section may contain a rippled wall, or rippled stream, or velocity jumps. Let us assume that this section is divided into many drift regions in tandem, with transition regions between successive sections as shown in Fig. 1(b). All of the even-numbered drift regions are the same, as are the odd ones and the odd-to-even transitions are the same, as are the even-to-odd, as indicated in Fig. 1(c).

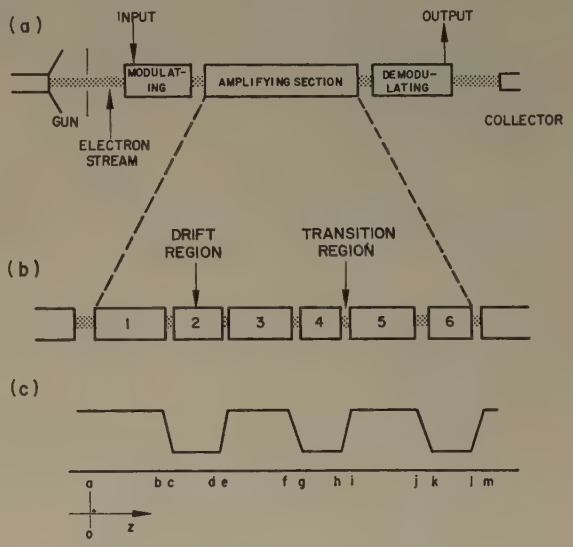


Fig. 1—Schematic representation of a typical rippled-wall, or rippled-stream, or velocity jump amplifier. In (a) the modulating sections could be cavities or helixes. In (b) the amplifying section is shown consisting of several quarter-plasma wavelength sections. In (c) the wall diameter, or stream diameter, or potential, is indicated for each section.

In the drift regions let us make the common assumptions of a stream of electrons constrained (by a very strong magnetic field) to flow in the z direction so that they are acted on effectively only by the z component of the electric field. Let the average current density and velocity be constant over the stream cross section $r<b$, forced to be so by positive ions which have very large mass and are neglected in the wave solution. As usual, only small signals are considered in which products of the ac quantities are neglected. In the transition regions, where changes in the wall or stream diameter or stream velocity occur, let us assume that the effective transition length is quite short compared to a quarter-plasma wavelength.

The problem of analysis is to express the modulation of the stream at the exit from the amplifying section in terms of the input modulation such that gain, equal

[7] Birdsall and Whinnery, *loc. cit.*, p. 320.

to the ratio of exit to input modulation, may be obtained. The technique of solution, following Tien,[6] is to obtain the ac current and velocity modulation at the end of each drift region using the space charge wave coefficients and to evaluate the transition effect using the Llewellyn[8] coefficients for an open-circuited gap.

The following notation is employed:

i, i_0 = ac, dc current densities ($i_0 < 0$, a function of z)
I, I_0 = ac, dc currents ($I_0 < 0$, not a function of z)
v, u = ac, dc velocities
V = dc potential with respect to cathode; $u = (-2\eta V)^{1/2}$
$\beta_e = \omega/u$ = stream wave number
$\lambda_e = 2\pi/\beta_e$ = stream wavelength
$\beta_p = \omega_p/u = [\eta i_0/\epsilon_0 u^3]^{1/2} = [\eta I_0/\pi b^2 \epsilon_0 u^3]^{1/2}$ plasma wave number
$\lambda_p = 2\pi/\beta_p$ plasma wavelength
p = plasma frequency (or wave number) reduction factor, a function of stream diameter b and wall diameter a and β_e.
$\omega_q = p\omega_p$, $\beta_q = p\beta_p$ } effective values
η = electron charge-to-mass ratio = -1.76×10^{11} coulomb/kg.
ϵ_0 = dielectric constant of vacuum = $(36\pi \times 10^9)^{-1}$ farad/m.

One obtains for the kth drift region, the current and velocity modulation at distance z from the input modulation, $I(0)$ and $v(0)$,

$$I(z, t) = [I(0) \cos (\beta_q z)_k - j g_k v(0) \sin (\beta_q z)_k] e^{j(\omega t - \beta_{ek} z)} \quad (1)$$

$$v(z, t) = \left[\frac{-j}{g_k} I(0) \sin (\beta_q z)_k + v(0) \cos (\beta_q z)_k\right] e^{j(\omega t - \beta_{ek} z)} \quad (2)$$

where

$$g_k = -\left[\frac{\omega}{\omega_q}\frac{I_0}{u}\right]_k, \text{ a positive quantity.}$$

Tien[6] noted that if I/v is defined as admittance and g taken as the characteristic admittance, then the expressions (1) and (2) [without $\exp(-j\beta_{ek}z)$] have the same form as the expressions for current and voltage of a lossless transmission system. Bloom and Peter[2] have actually built transmission line analogs in which the standing wave measured along the line represented the amplitude pattern of the space charge wave along the beam.

In the simplest case of interest, the drift regions are all a quarter-plasma wavelength long, $\beta_q z = \pi/2$ for all sections, such that (1) and (2) reduce to:

$$I = -jg_k v(0) \quad (3)$$

$$v = \frac{-jI(0)}{g_k}. \quad (4)$$

[8] F. B. Llewellyn, "Electron Inertia Effects," Cambridge University Press, Cambridge, Eng.; 1943.

(Actually these results are true, except for sign, for $\beta_q z = n\pi/2$ where n is any odd integer; also, some sections may be, say $\pi/2$ long, others $3\pi/2$ long.)

For the simple gap assumed, I is continuous and the product uv is continuous.[6,8] Hence, after two drift regions and two gaps one obtains

$$\left|\frac{I(e)}{I(a)}\right| = \frac{g_2}{g_1}\frac{u_1}{u_2} = G, \quad (5)$$

$$\left|\frac{v(e)}{v(a)}\right| = \frac{g_1}{g_2}\frac{u_2}{u_1} = \frac{1}{G}, \quad (6)$$

and, in general, after $2N$ drift rerions and $2N$ gaps,

$$\left|\frac{I}{I(a)}\right| = G^N \qquad N = 1, 2, 3, \cdots \quad (7)$$

$$\left|\frac{v}{v(a)}\right| = G^{-N}. \quad (8)$$

Gain parameter G may be expanded into separate factors to make clear the various gain mechanisms.

$$G = \left(\frac{V_1}{V_2}\right)^{3/4}\left(\frac{b_2}{b_1}\right)\left(\frac{p_1}{p_2}\right). \quad (9)$$

The first factor may be thought of as the familiar velocity jump term, the second factor as the rippled stream term, and the third factor as the rippled wall term. However, it should be noted that the last factor may be different from unity for both a velocity jump amplifier with smooth walls, and a rippled stream amplifier in which only beam diameter varies.

It is of interest to compare the rate of amplification with that of other types of tubes which have waves increasing exponentially with distance. One sees, on simple examination, that the gain here also is exponential with distance. Let the distance from a to e, or b to y, or e to i, etc., be defined as

$$\frac{\lambda_{qm}}{2} = \frac{\lambda_{q1}}{4} + \frac{\lambda_{q2}}{4} \quad \text{and} \quad \beta_{qm} = \frac{2\pi}{\lambda_{qm}}. \quad (10)$$

Then, at distances L from plane a, $z = 0$, where

$$L = N\left(\frac{\lambda_{qm}}{2}\right), \quad N = 1, 2, 3, \cdots, \quad (11)$$

one finds

$$\left|\frac{I(L)}{I(a)}\right| = [G]^N = [G]^{\beta_{qm}L/\pi}. \quad (12)$$

This growth is shown in Fig. 2 (following page). In decibels the gain is

$$\text{gain} = 10 \log_{10}\left|\frac{I(L)}{I(a)}\right|^2 = \frac{20}{\pi}\beta_{qm}L \log_{10} G \text{ db}. \quad (13)$$

Gain per plasma wavelength ($\beta_{qm}L = 2\pi$) is then

$$\text{gain} = 40 \log_{10} G \text{ db}. \quad (14)$$

This figure may be compared to gains of nonpropagating, but growing-wave, tubes,[1]

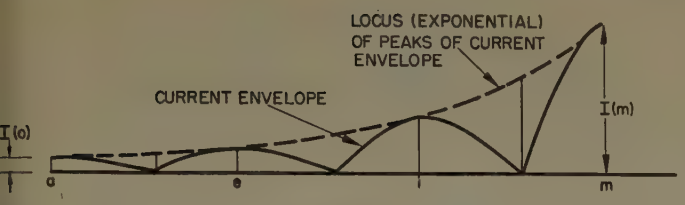


Fig. 2—Envelope of the current modulation as a function of distance. The peaks are separated by one-half plasma wavelength and follow an exponential curve.

$$\text{gain} = 54.6q \text{ db}. \tag{15}$$

As examples, the gain-factor $q=0.351$ for a purely resistive wall, so $54.6q=19.2$ db; $p_1/p_2\cong 2$ for a possible rippled-wall tube, so $40 \log_{10} G=12$ db. Comparison to traveling-wave tube rate of gain is not straightforward as such gain is proportional to $(I_0)^{1/3}$ or $(I_0)^{1/4}$, here to $(I_0)^{1/2}$.

In general, $(\beta_q z)_k$ is a function of frequency and operating parameters of the tube and hence may vary from the optimum value, $\pi/2$, with changes in these factors. For example, consider that all

$$(\beta_q z)_k = \pi/2 + \delta. \tag{16}$$

In order to obtain some idea of the behavior of gain as δ varies from zero, consider the simple case of current modulation only and $G \gg 1$. For N sections one obtains

$$I(N)/I(a) = (G \cos^2 \delta - \sin^2 \delta)^N, \tag{17}$$

and for δ small

$$\cong G^N(1 - \delta^2/2)^{2N} \cong G^N(1 - N\delta^2). \tag{18}$$

Thus the gain variation is parabolic around the maximum value of G^N, being narrower as N increases; this behavior is like that of a constant gain amplifier followed by a maximally flat bandpass filter. For example, $N=5$, gain is 0.707 of G^N for $\delta\cong\pm 15$ degrees (half-power points). $N\delta^2$ is then a measure of bandwidth or of criticalness of the operating parameters.

Evaluation of Reduction Factor p

The simple theoretical description is applicable to streams of arbitrary cross section which change shape only in the transition regions. The theory is completed by evaluation of the plasma frequency reduction factor, p. This factor accounts for the finiteness of the stream and the effect of charges on distant or nearby electrodes. If the stream is broad, compared to the distance between bunches (λ_s), then the stream behaves much as if it were infinitely broad, $p\cong 1$. If the stream is not broad, then p can be quite small and very much dependent on the nearness of electrodes.

In order to calculate G, it is necessary to know p as a function of the stream and circuit parameters. The simplest configuration is that of a circular solid stream confined by a very strong magnetic field to flow through a metal drift tube; this model is well known,[1,3,4] leading to p as shown in Fig. 3(a) and (b). The parameters are

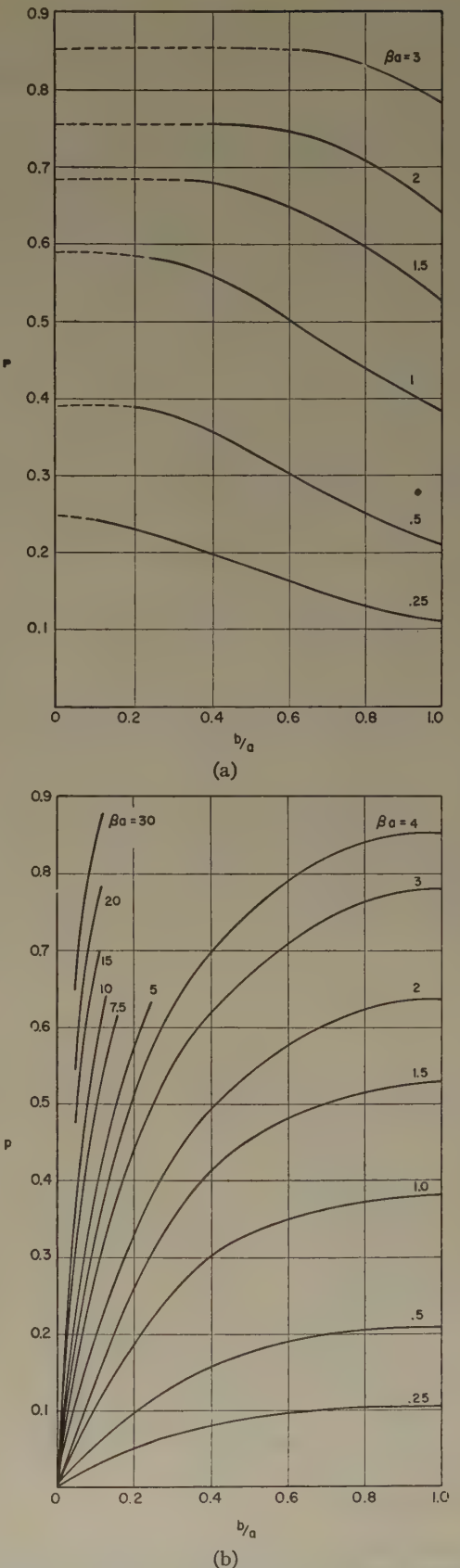


Fig. 3—Plasma frequency reduction factor p for a solid stream of radius b in confined flow in a perfectly conducting drift tube of radius a. The fundamental mode is assumed in which the electric field has one radial variation ($E_z=0$ only at the wall) and no angular variation. (a) Parameter βb, useful in rippled-wall gain calculation (b=constant). (b) Parameter βa, useful in rippled-stream gain calculation (a=constant).

βb and βa which depend on $p(\beta=\beta_e\pm\beta_r p)$ such that a trial and error process is necessary to obtain p; however, for most work $\beta\cong\beta_e$, requiring but one trial for negligible error.

Several other configurations are calculable, but not included here. Thin hollow, or Brillouin flow, or ion neutralized streams, all three having ac current concentrated at the outer edge of the stream, have reduction factors that go to zero as $b/a\rightarrow 1$. Hence, the ratio, p_1/p_2, can be made much larger than for solid confined flow streams leading to larger amplification in the rippled wall case.

Typical Tubes and Some Design Curves

For typical amplifier applications, where the mechanism under discussion is to provide the bulk of the gain, the structure of this class of tubes is essentially as shown in Fig. 1(a). For other applications, such as in mixers and detectors, the arrangement would be suitably modified.

Of course, attenuation also may be obtained as is useful in noise reduction. In this application, the deamplifying section follows the electron gun and precedes the interaction region. For example, in Fig. 1(c), noise velocity modulation at $z=a$ is reduced at e or i or m for $G>1$, and similarly for noise current modulation, with $G<1$.

In a typical tube, all three factors of G differ from unity. One factor predominates and lends its name to the mechanism of gain; unfortunately, the other two may, and generally do, diminish gain. Let us consider a rudimentary model of each type and some possible refinements.

Velocity-jump amplifier

The predominant factor in G is $(V_1/V_2)^{3/4}$. Due to the lens action of the jumps, the stream diameter may vary periodically; b_2/b_1 approaches unity only in the limit of a very large confining magnetic field. (The lens action, of course, might be used to aid in or even replace magnetic focusing.) Also, if all the drift tube diameters are the same, then reference to Fig. 3 shows that $p_1/p_2<1$, detracting from gain. This effect could be reduced by making the low-voltage drift-tube diameter smaller than that of the high-voltage tube. A typical section might be as given in Fig. 4, which shows two refinements, changes in diameter of stream and drift tube.

No design curves are included. Tien[6] indicated (but did not publish) that "detailed design procedures . . . have been obtained."

Rippled wall amplifier

The predominant factor in G is of course p_1/p_2. In the simplest model, Fig. 5, the other factors are unity, and design curves are easily obtained.

Of primary interest is optimum design for maximum

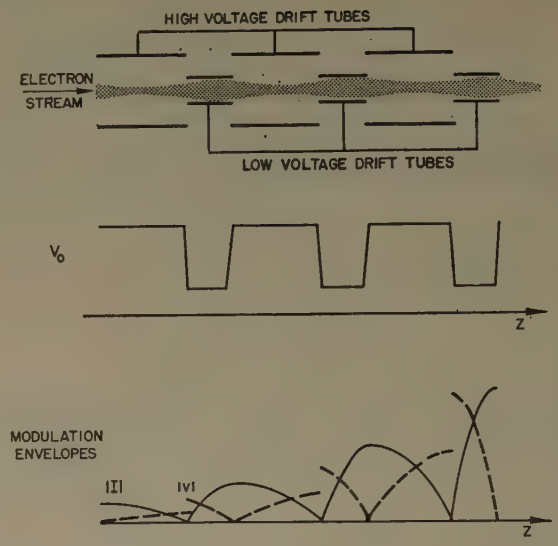


Fig. 4—Velocity jump amplifying section with added refinements of a rippled wall and rippled stream for optimizing gain. The dc potential on the drift tubes is shown and the ac current and velocity modulation envelopes.

rate of gain. From (14), one sees that increasing $G(=p_1/p_2)$, increases gain per mean plasma wavelength. From Fig. 3, it may be shown that p_1/p_2 increases as βb decreases. However, at the same time, p_1 and p_2 decrease, which decreases β_{qm}. The net effect is to decrease gain per meter as βb decreases below a certain value. What is desired, then, is a knowledge of the rate of gain per meter such that optimum design is evident. This is obtainable directly from a revised form of (13), using (10):

$$\text{gain} = \frac{20}{\pi}\beta_{qm}L\log_{10}G\ \text{db} \tag{13}$$

$$= \frac{40}{\pi}\beta_p L\left(\frac{1}{p_1}+\frac{1}{p_2}\right)^{-1}\log_{10}\frac{p_1}{p_2}\ \text{db} \tag{19}$$

$$= \frac{40}{\pi}\beta_p F(p_1, p_2)\ \text{db/meter}. \tag{20}$$

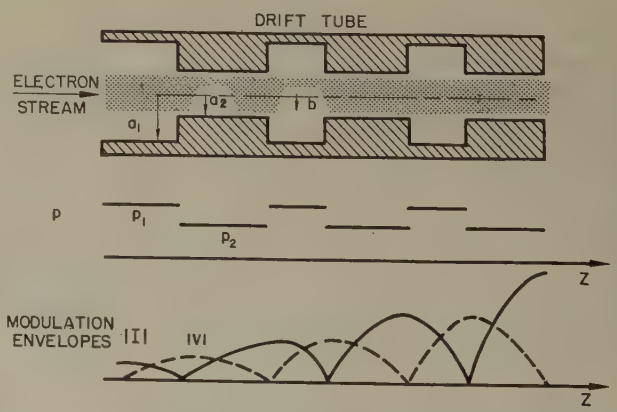


Fig. 5—Rippled-wall amplifying section with simple drift tube. Also shown are the reduction factors $p_{1,2}$ and ac current and velocity modulation envelopes.

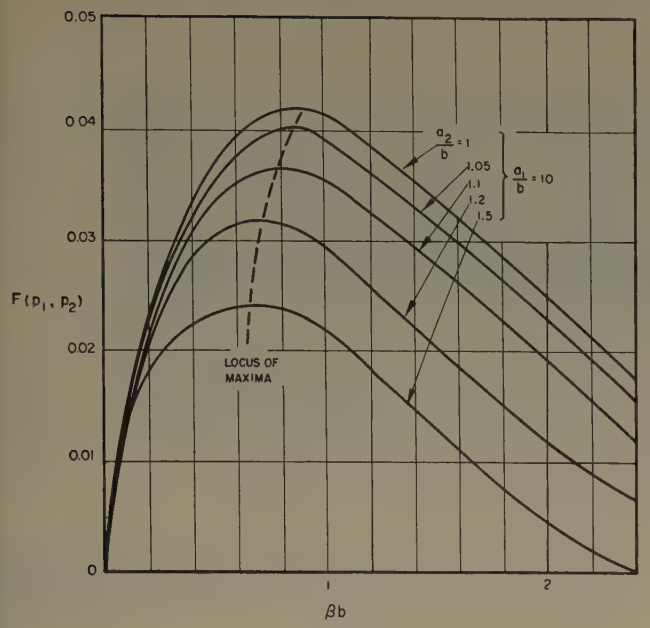


Fig. 6—Gain parameter $F(p_1, p_2)$ of rippled-wall amplifier (solid stream, confined flow, fundamental mode) as a function of βb and several drift minimum wall radii (a_2). The locus of maxima (design points for highest gain/meter) is shown, varying between βb of 0.7 and 0.9.

The function, $F(p_1, p_2)$, is obtained from Fig. 3 and is plotted in Fig. 6 for a solid stream with several minimum wall diameters, and one maximum diameter (which, though chosen large, could have been one-half or one-third as large with little change in F). Obviously, optimum design is with βb in the range 0.7 to 0.9. Maximum obtainable gain, using $a_2/b = 1$, is:

$$\text{maximum gain} = \frac{40}{\pi} \times 0.042\beta_p = 0.535\beta_p \text{ db/meter.} \quad (21)$$

For one-dimensional models of resistive, medium, or double stream tubes, maximum gain is about $3\beta_p$ db/meter, roughly six times the above rate of gain; for actually realizable models, the rate is a fraction of $3\beta_p$. Even so, rippled wall amplification (in solid stream, confined flow) cannot be considered large. The selling points of this type are the utter simplicity of circuit construction and the ability to use Brillouin flow, as contrasted with the velocity jump tube.

It should be observed that the curves of F versus βb are not curves of gain versus frequency or inverse velocity for a given tube. Rather, F is a locus of maxima of gains for varying p_1/p_2, hence, varying lengths of drift sections, as is possible only in different tubes.

Rippled stream amplifier

In the simplest model, Fig. 7, the drift tube is of constant diameter; the predominant factor in G should be b_2/b_1. Analogous to the velocity-jump amplifier (where V_1/V_2 can be made very large), G can be made quite large (by increasing b_2/b_1); this is in contrast to the

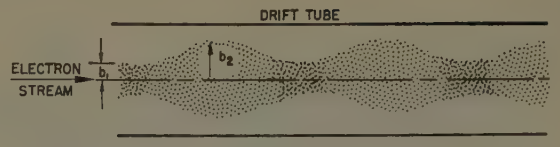


Fig. 7—Rippled-stream amplifying section with smooth-walled drift tube. The rippling is caused by magnetic or electric lenses not shown.

rippled wall amplifier, using a simple drift tube, where p_1/p_2 cannot be made very large. However, it should be noted that p_1/p_2 is less than unity if b_2/b_1 is greater than unity; consequently, the maximum stream diameter must be large in electronic wavelengths if the latter effect is not to cancel the gain.

The effect of the wall is detrimental even though the stream is nearest the wall when bunching. The reason for this is that when the stream diameter is large, compared with the distance between bunches (βb large), the image charges can reduce the field between bunches very little (p large), but oppositely for a comparatively small diameter. Hence, the forces due to image charges work in opposition to the gain mechanism.

Electron streams ordinarily occur rippled and, indeed, are difficult to keep from rippling. For small radial amplitude, the rippling may be considered sinusoidal with z, stationary in space, or with t, an electron taking time, τ, to go from one peak to the next. That is, in time τ, the stream has contracted then expanded back to its original diameter, completing one full cycle of radial motion. If, during this excursion, the phase of the modulation is such that the stream bunches while expanded and debunches while contracted, then gain in current will result. Thus, the bunching period should be 2τ.

One simple example is perturbed Brillouin flow in which $2\pi/\tau$ is ω_p (see, for example, Pierce[9]). As argued above, $2\omega_q$ should equal ω_p or $p = 0.5$ ($p_1 \cong p_2$). This value of p is obtainable, occurring at a frequency dependent on the ratio $b/a (b_1 \cong b_2)$. Rigrod[10] shows that $p = 0.5$ at $\beta b = 1.4$ for $b/a = 0.6$, smaller βb for smaller b/a. (Note that p for Brillouin flow is not the same as for confined flow as shown in Fig. 3.) As some rippling usually accompanies what is designed to be perfect Bril louin flow, some gain is to be expected always, desired or not, though small for small rippling.

In confined flow, the stream always is rippled with approximate radian frequency

$$\frac{2\pi}{\tau} \cong \omega_{cp} = \omega_p \sqrt{\frac{\overline{\omega_c^2}}{\omega_p^2} - 1} \quad (\omega_c = -\eta B_z) \quad (22)$$

[9] J. R. Pierce, "Theory and Design of Electron Beams," D. Van Nostrand Co., Inc., New York, N. Y., pp. 155–157; 1949.

[10] W. W. Rigrod and J. A. Lewis, "Wave propagation along a magnetically-focused cylindrical beam," *Bell Sys. Tech. Jour.*, vol. 33, pp. 399–416; March, 1954.

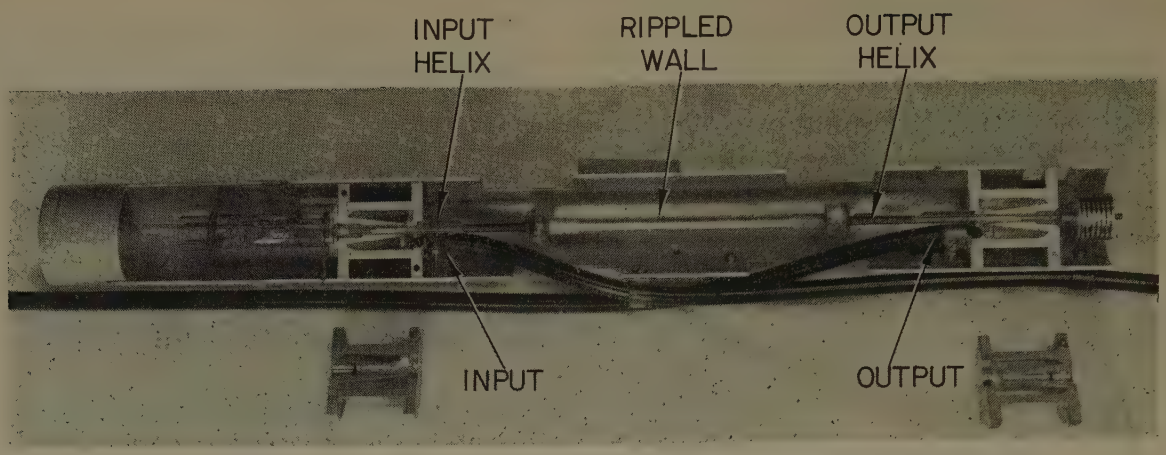


Fig. 8—Rippled-wall amplifier with helix modulating sections for use in the 1 to 4 kmc range.

such that for gain

$$p \cong \frac{1}{2}\sqrt{\frac{\omega_c^2}{\omega_p^2}} - 1 \cdot \tag{23}$$

Again for this value there is a frequency, dependent on b/a and ω_c/ω_p, found from Fig. 3. Here, note that with a drift-tube circuit, gain is unobtainable for $\omega_c > 2.24\ \omega_p$ because this requires $p > 1$, impossible for this circuit (though possible for others, such as those with an inductive wall admittance).

Rate of gain is calculable from (13), for the stream with sudden expansions and contractions, and design criteria, may be obtained. Of course, actual streams ordinarily would have more or less sinusoidal boundaries so that this calculation is quite approximate. As in the previous section, the expression for gain may be re-cast to more useful form

$$\text{gain} = \frac{40}{\pi} L[(\beta_{p1}p_1)^{-1} + (\beta_{p2}p_2)^{-1}]^{-1} \log_{10} \frac{b_2}{b_1}\frac{p_1}{p_2} \text{ db.} \tag{24}$$

One consideration in design is to avoid approaching limiting current[11] which will be determined by the small diameter section. This limit can be given in terms of $\beta_{p1}b_1$ and b_1/a (e.g., for confined flow, $b_1/a = 0.1$, one finds limiting current is given by $\beta_{p1}b_1 = 0.39$). For this reason β_{p1} is factored out:

$$\text{gain} = \frac{40}{\pi} \beta_{p1}p_1[1 + G]^{-1} \log_{10} G \text{ db} \tag{25}$$

$$\text{gain} = \frac{40L}{\pi} \beta_{p1}p_1 F(G) \text{ db.} \tag{26}$$

$F(G)$ is a simple function with a maximum value of 0.121 at $G = 3.58$. However, in the range $2 < G < 8$, one finds $F(G) > 0.1$. With so broad a curve, there is a very wide range of values of βa, b_2/p_1 and p_1/p_2 which lead to large $F(G)$, such that curves showing the particular values leading to $G = 3.58$ could be misleading. Sample gain calculations show that highest gain is obtained for:

(a) large values of b_2/b_1 (to make G large),
(b) large values of βa (to make p_1 as large as possible),
(c) $b_2 \cong a$ (so that, for a given b_2/b_1, p_1/p_2 is largest).

Although (a) and (b) are true for the model used, in actual design they must be applied with caution and not taken to extremes for several reasons. Concerning (a), very large expansion would require a large fraction of $\lambda_q/4$ for "gap" length, which invalidates the analysis. Concerning (b), at very large βa higher order space charge modes would be excited to appreciable amplitude either initially or in the expansion "gaps;" these modes are neglected in the analysis.

The maximum gain obtainable is:

$$\text{maximum gain} = \frac{40}{\pi} \beta_{p1}p_1(0.121) = 1.54\beta_{p1}p_1 \text{ db/meter.} \tag{27}$$

Although β_{p1} may be very large (being measured in the smaller diameter section), p_1 may be quite small; hence, the product, $\beta_{p1}p_1$, may have about the same value as β_p of a comparable rippled wall amplifier. If this is true, then the ripped stream gain is three times as large.

Although axial variation in charge density is here associated with axial change in stream diameter, this implies radially uniform charge density. However, axial variation may exist in other models. One such case is that of a solid stream with nearly smooth outer diameter, but with inner electrons having periodic radial motion in an organized fashion. This behavior is possible, for example, in (imperfect) Brillouin flow for a stream in which charge density has been caused to vary with radius. G. R. Brewer has noted the possibility of this type of flow and found, for parabolic charge density variation, that the inner electrons can perform simple harmonic motion radially while the outermost electrons have linear trajectories. In experimenting, then, measurement of the rippling of the outer diameter is insufficient to determine whether or not the stream has periodic charge density variations. Perhaps this "inner rippling" may explain why gain has been observed by some experimenters where outer rippling was small.

[11] Pierce, *op. cit.*, pp. 161–164.

Description of Experiment With Rippled Wall Amplifer

One experimental tube was constructed and tested for the sole purpose of demonstrating the existence of rippled wall amplification. The tube design was not optimum, but made to fit existing parts in our laboratory. The model consisted of a rippled-wall drift tube of six sections, each a quarter-plasma wavelength long, between two helixes which were used to modulate and demodulate the stream. Fig. 8 (opposite page) is a photograph of the tube in its cradle.

Indication of rippled wall gain was found in two ways. First, the gain was compared with that obtained using a smooth-wall drift tube in an earlier experiment. This comparison was made over a wide range of currents and frequencies and showed an increase due to the rippled wall (about 6 to 9 db) over that from the drift tube. However, a more clear-cut indication was to observe gain versus rippled wall potential at several values of frequency and stream current. This variation was observed on an oscilloscope. Typical traces were similar to the point by point measurements shown in Fig. 9.

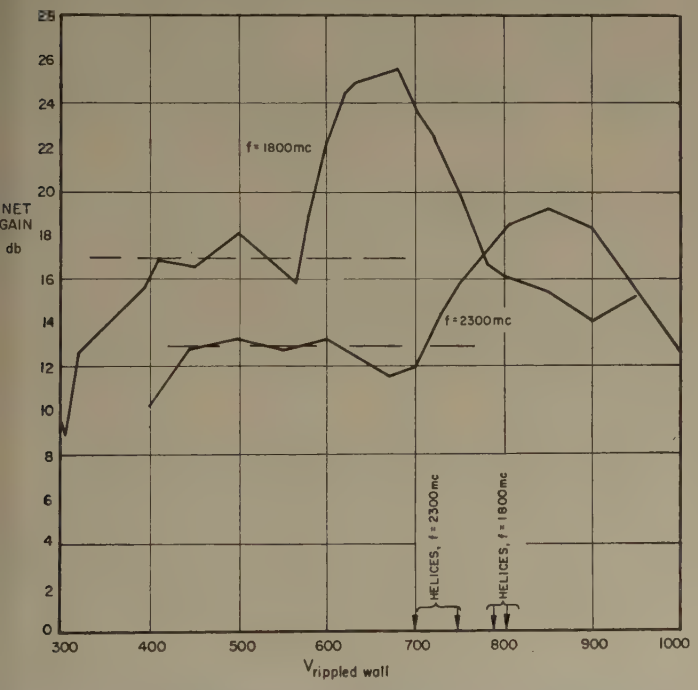


Fig. 9—Net gain of rippled-wall amplifier at 1.8 kmc and 2.3 kmc as a function of the drift-tube potential. The helix potentials are indicated. Measurement was made at separated points as shown; actual gain varied in a smoother fashion. Dashed lines show approximate gain obtained with a smooth-wall drift tube.

As the helixes were held at constant potential (about 700 to 800 volts), some velocity jump gain might be expected, especially at potentials much larger or smaller than helix potential. However, there seems to be little indication of these effects. The conclusion is that the peak observed (which was near helix potential) was due solely to rippled wall amplification.

The dimensions of the drift-tube sections were obtained using confined flow reduction factors. The original design parameters are listed in Table I. After the tube was tested, it appeared desirable to see if the observed frequency, voltage, and current characteristic for optimum gain, agree with calculation. This check requires a brief addition to theory.

TABLE I
Design Parameters

Cathode diameter	Drift Tube diameters	Drift Tube lengths	p_1/p_2	$\beta_e b$	p_2, p_1	Gain in six $\lambda_q/4$ sections
0.050″	0.0625″ 0.200″	0.8″ 0.6″	1.33	1.0	0.440 0.585	$\binom{6}{-}{4}$ 40 $\log_{10} p_1/p_2$ = 7.5 db

Gun design was parallel flow Pierce type.
$\beta_e b$ was chosen to give $p_1/p_2 = 1.33$ for b/a's of $0.050/0.0625 = 0.8$, $0.050/0.200 = 0.25$.

For a given tube the λ_{q1}'s must be near $\pi/4$ in order to obtain constructive interference, that is, optimum gain. This requirement determines the f, V_0, I_0 characteristic:

$$\beta_p p_1 l_1 = \beta_p p_2 l_2 \cong \pi/4. \tag{28}$$

For very small $\beta_e b$, $p \sim \beta_e b$; this relation is taken to hold approximately in the region of interest. (Thus $p_1 \cong 0.585\ \beta_e b$, $p_2 \cong 0.440\ \beta_e b$. This implies p_1/p_2 constant, which is a good assumption for small variations in $\beta_e b$.) One then obtains:

$$(f)(I_0/V_0^{5/2})^{1/2} = \text{constant}. \tag{29}$$

At three currents (5, 10, and 15 ma) and at three frequencies (1.8, 2.3 and 2.8 kmc) the behavior was found to follow this law very closely, the experimental average for the constant being 0.649×10^5 with maximum deviation being ± 6 per cent. However, the constant as calculated from the values of Table I is 0.476×10^5. For a smaller stream (0.040 inch diameter) the calculated constant is 0.368×10^5; for a larger stream (0.0625 inch diameter, largest possible), it is 0.595×10^5. Thus, although the functional law seems valid, the check between measured and calculated value of the constant is only fair.

An explanation of this discrepancy is that the stream was not in confined flow but in some other regime for which the reduction factors were smaller. Either ion-neutralized flow or Brillouin flow are possibilities. The former seems improbable as trapping potentials were present only in some regions of operation. The latter is a better candidate as the cathode was placed near the entrance to the solenoid where the flux density was 56 per cent of that in the drift-tube regions. The author sought best current transmission, forgetful of the ac-

companying change in ac behavior. Thus the stream was at least partly magnetically focused. In either case the reduction factors are the same and are substantially smaller than for confined flow.

Using these reduction factors, the experimental constant is found to hold, assuming a 0.048 inch diameter stream, which is well within reason.

Although the tube was operated under conditions different from the original design, the author is confident that rippled wall amplification was the main cause of the observed gain.

ACKNOWLEDGMENT

The idea of rippled wall amplification was shared with D. A. Watkins. Discussions with A. V. Haeff and J. R. Whinnery, concerning the rippled stream work, were most valuable. The tube design was worked out jointly with G. R. Brewer. The actual tube construction was carried through by T. E. Gorman and R. T. Morris. Their share in this work was substantial and is gratefully acknowledged.

Most of the theory and experimental design reported was completed before the end of 1952.

Stability Considerations in the Parameter Measurements of VHF Point-Contact Transistors*

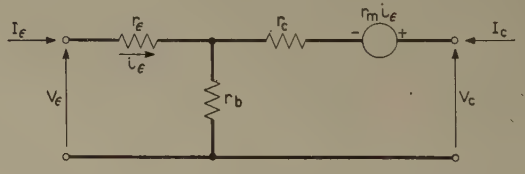
DONALD E. THOMAS†, SENIOR MEMBER, IRE

Summary—**This paper considers stability problems encountered in measuring parameters of vhf point contact transistors. It develops the high frequency modification of the low frequency requirement for stability. It then considers instability which may be caused by negative resistance collector regions. This type of instability is responsible for breaks or discontinuities which have been observed in the static characteristics of some point contact transistors. Simple stabilized circuits for static characteristic curve tracers, using either oscilloscopic or X-Y recorder presentation, are given.**

INTRODUCTION

ONE OF THE EARLIEST THINGS learned about a point-contact transistor was that is was short-circuit unstable. In order to avoid this instability it was natural to use high-impedance constant-current biasing sources for both the emitter and the collector of the transistor, and to use high-impedance, constant-current signal generators in measuring the transistor. This naturally led to a description of the transistor in terms of its open-circuit impedances. The equivalent circuit resulting from this approach to the problem of describing the transistor is given in Fig. 1, along with the definitions of the individual open-circuit resistances from which the equivalent circuit is determined.

Ryder and Kircher pointed out very early that, although this description of the transistor was convenient, it was neither final nor unique; and, in fact, as soon as the higher frequency performance of transistors became of interest, one of the other possible descriptions became more convenient.[1] This description involved the substitution of a current generator, αi_e, for the voltage generator, $r_m i_e$, in the collector circuit in Fig. 2, p. 1637. The definition of the dimensionless constant alpha is also given in the figure. Now note that one of the elements of the equivalent circuit is defined in terms of a



$$r_e = r_{11} - r_{12} \qquad r_{11} = \frac{\partial V_e}{\partial I_e} \qquad I_c = \text{constant}$$

$$r_b = r_{12} \qquad r_{12} = \frac{\partial V_e}{\partial I_c} \qquad I_e = \text{constant}$$

$$r_c = r_{22} - r_{12} \approx r_{22} \qquad r_{22} = \frac{\partial V_c}{\partial I_c} \qquad I_e = \text{constant}$$

$$r_m = r_{21} - r_{12} \approx r_{21} \qquad r_{21} = \frac{\partial V_c}{\partial I_e} \qquad I_c = \text{constant}$$

Fig. 1—Transistor equivalent circuit in terms of open-circuit impedances.

constant-collector voltage or, in other words, a short-circuit collector termination. That this is permissible without encountering instability can be seen from an examination of the low frequency criterion of stability[1] which is given below:

$$\frac{r_m}{R_c} < 1 + \frac{R_e}{R_b} + \frac{R_e}{R_c}. \qquad (1)$$

The capital R's are the total resistances, internal and external, in the leads indicated by the subscripts. The stability requirement of (1) shows that, even though the collector is short-circuited, stability can be obtained by making the external resistance in the emitter lead sufficiently high and vice-versa.

* Decimal classification: R282.12. Original manuscript received by the IRE, July 7, 1954.

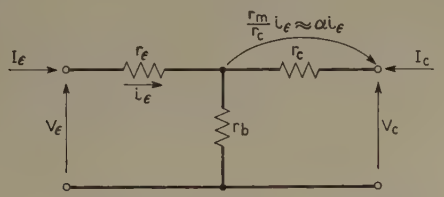
† Bell Telephone Labs., Murray Hill, N. J.

[1] R. M. Ryder and R. J. Kircher, "Some circuit aspects of the transistor," *Bell Sys. Tech. Jour.*, vol. 28; pp. 367–400; July 1949.

Early Transistor Measurement Equipment

Practically all early transistor measurement equipment was designed to measure the four open-circuit and single combination open-circuit short-circuit parameters defined in Figs. 1 and 2. Since early models of transistors had alpha frequency cutoffs which were relatively low, it was not necessary to give important consideration to lead impedances and parasitic shunt capacitances. Even in the short-circuit collector measurement of alpha, the collector was short-circuited by design only at the measurement frequency. High-impedance, constant-current[2] dc collector bias was still used, and measurements rather than stability considerations determined the collector terminating impedance at frequencies differing from the measuring frequency.

When point-contact transistors having alpha cutoff frequencies in the vhf region were developed, they proved to be unstable in practically all of these early designs of measuring equipment. A complete re-evaluation of stability requirements was therefore necessary, in order to modify old and to construct new measuring equipment suitable for characterizing these higher frequency units.



$$\alpha = -\frac{\partial I_c}{\partial I_e}\Big]_{V_c=\text{constant}}$$

$$\alpha = \frac{r_{21}}{r_{22}} = \frac{r_m + r_b}{r_c + r_b} \approx \frac{r_m}{r_c}$$

Fig. 2—Equivalent circuit with alpha current generator.

Re-Evaluation of Stability Requirements

Consideration was first given to the necessary modification of the low-frequency requirement for stability of transistors given in (1), which took into account parasitic capacitances and the frequency characteristic of the alpha current gain of the transistor. The circuit chosen for this evaluation is given in (a) of Fig. 3. A short-circuit collector was chosen since, with proper emitter-to-collector shielding to keep emitter-to-collector capacitance small, this is the most severe condition. C then becomes the sum of the parasitic emitter-to-collector and emitter-to-ground capacitances of the transistor, plus any residual capacitance due to external wiring, sockets, etc. The equivalent circuit of the circuit in (a) of Fig. 3 is given in (b) of Fig. 3. The equations relating the voltages and currents of the circuit of Fig. 3(b) are:

$$\begin{aligned}
\left(r_g - j\frac{1}{\omega C}\right) i_1 \qquad & + j\frac{1}{\omega C}\, i_2 \qquad = 0 \\
+ j\frac{1}{\omega C}\, i_1 + \left(r_e + r_b - j\frac{1}{\omega C}\right) i_2 \qquad & - r_b i_3 = 0 \qquad (2) \\
& -(r_b + r_m) i_2 + (r_c + r_b) i_3 = 0.
\end{aligned}$$

If we replace r_m by its approximate equivalent, αr_c, we can take into account the frequency dependence of the current gain of the transistor by using the usual expression for α as a complex function of frequency[3] which is

$$\alpha = \frac{\alpha_0}{1 + j\dfrac{f}{f_c}}, \qquad (3)$$

where:

α_0 = the low-frequency magnitude of alpha; and
f_c = the frequency at which the magnitude of alpha is down by 3 db from its low-frequency value.

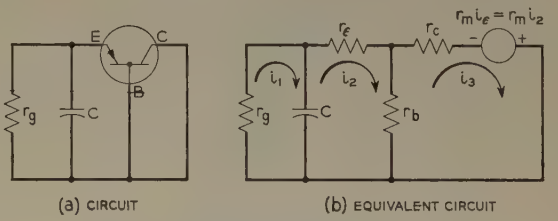


Fig. 3—Potential relaxation oscillator circuit. (a) Circuit. (b) Equivalent circuit.

If, after making the above substitutions, we then eliminate the complex operator j by substituting the complex variable p for $j2\pi f$, the circuit determinant for the circuit of Fig. 3(b) as determined from equations (2) is given by:

$$\Delta = \begin{vmatrix} r_g + \dfrac{1}{pC} & -\dfrac{1}{pC} & 0 \\ -\dfrac{1}{pC} & r_e + r_b + \dfrac{1}{pC} & -r_b \\ 0 & -r_b - \dfrac{\alpha_0 r_c}{1 + \dfrac{p}{2\pi f_c}} & r_c + r_b \end{vmatrix}$$

$$= \frac{2\pi f_c}{pC(p + 2\pi f_c)} \left\{ \frac{r_g C}{2\pi f_c} \left[r_e r_c + r_c r_b + r_e r_b \right] p^2 \right.$$

$$+ \left[r_g C (r_e r_c + r_c r_b + r_e r_b) \right.$$

$$\left. + \frac{1}{2\pi f_c} (r_c r_e + r_e r_b + r_b r_c) \right.$$

[2] An ideal constant-current generator would have infinite impedance and an ideal constant-voltage generator would have zero impedance. However, to be effectively constant-current a generator need have only an impedance high enough relative to the impedance of the load it is driving that the load impedance has only a second order effect on the current which flows in the circuit. Similarly, a constant-voltage generator need have only an impedance low enough relative to its load that the load impedance has only a second-order effect on the voltage across the circuit. In all discussions to follow, these latter definitions will be descriptive of what is meant when a constant-current or constant-voltage generator is referred to.

[3] D. E. Thomas, "Transistor amplifier—cutoff frequency," Proc. I.R.E., vol. 40, pp. 1481–1483; November, 1952.

$$+\frac{r_g}{2\pi f_c}(r_e+r_b)\Big]p$$

$$+[(r_e r_g+r_g r_b+r_b r_e+r_e r_c+r_c r_b-\alpha_0 r_c r_b)]\Big\}. \quad (4)$$

Now, it is required for stability that the determinant of the circuit as given by (4) have no zeros in the right half of the p plane.[4] The zeros of (4) are given by the roots of the second-degree polynomial in p between the $\{\}$ braces in (4). Routh's criterion tells us that each of the roots of a second-degree polynomial will have negative real parts (i.e., will be in the left half of the p plane) if all of the coefficients of the polynomial including the p^0 coefficient, or constant term, are greater than zero.[5] Under the assumed conditions, the p^2 coefficient of the polynomial will always be greater than zero. It is therefore only required that the circuit elements be such that the p coefficient and the constant term of the polynomial be greater than zero if we are to have no roots of (4) in the right half of the p plane and therefore be assured of circuit stability. This will be true if the following two conditions, which become the requirements for circuit stability, are met:

$$\alpha_0 < 1+r_e\left(\frac{1}{r_b}+\frac{1}{r_c}\right)+\frac{1}{2\pi f_c r_g C}\left[1+(r_e+r_g)\left(\frac{1}{r_b}+\frac{1}{r_c}\right)\right]$$

and

$$\alpha_0 < 1+(r_e+r_g)\left(\frac{1}{r_b}+\frac{1}{r_c}\right). \quad (5)$$

If we assume $r_g \gg r_e$, $r_g \gg r_b$, and $r_c \gg r_b$, then the two requirements for stability become:

$$\alpha_0 < 1+\frac{r_e}{r_b}+\frac{1}{2\pi f_c r_b C} \quad (6)$$

and

$$\alpha_0 < 1+\frac{r_g}{r_b}.$$

Since we have assumed $r_g \gg r_b$, and since α_0 is usually of the order of magnitude of three or less throughout most of the operating region of point-contact transistors, the second condition is satisfied. However, since r_e/r_b is usually less than unity, the first requirement of (6) places a minimum permissible value on $1/(2\pi f_c r_b C)$, below which the circuit will be unstable for transistors having an α_0 of two or above. This places a maximum permissible value on the product of $f_c r_b C$ which cannot be exceeded if stability is to be assured. In order to get alpha cutoff frequencies of point-contact transistors in the vhf region, it is necessary to go to very close point spacings. Unfortunately, base resistance also increases rapidly as point spacing is decreased. Therefore, although germanium of lower resistivity is used to counteract this tendency of base resistance to rise, the base resistances of vhf point-contact transistors tend to be somewhat higher than those of units with lower cutoff frequencies. Thus, as f_c, the alpha cutoff frequency, is increased, r_b also increases. It is not surprising, therefore, that in many circuits designed when alpha cutoff frequencies were relatively low, the $f_c r_b C$ product for newly developed vhf point-contact transistors exceeded the maximum value permissible for stability as determined by (6), and that, therefore, the transistors oscillated in these measuring circuits.

The problem proved to be soluble by a simple circuit modification. Since constant-current emitter bias and signal sources are used to the emitter in each of the point-contact measurements so far defined, a resistance, large compared to r_e and r_b, is placed in the emitter input circuit. The physical position of this resistance is as near as possible to the emitter terminal of the transistor in order to assure that the capacitance C is the minimum possible as determined by the parasitic capacitance of the transistor itself. This resistance is shown in (a) of Fig. 8 by the resistance R_{e1}. All other emitter-circuit connections with one exception are then made on the V_{e1} end of R_{e1} and are therefore isolated from the emitter.

As shown by the definitions in Figs. 1 and 2, it is necessary to measure emitter potential in making parameter measurements. In order to avoid the error due to $I_e R_e$ [see (a) of Fig. 8] which would occur if the potential were taken at V_{e1}, a second resistance, R_{e1}', is connected directly to the emitter terminal to produce a potential terminal, V_e. Since this is a potential terminal, the resistance, R_{e1}', will usually be negligible or require only a small correction to the emitter-potential calibration. The resistances, R_{e1} and R_{e1}', in parallel now correspond to the resistance, r_g, of the stability solution of (5) and must therefore be large, compared to both r_e and r_b of the transistor, if the stability requirements of (6) are to be valid. The above solution to the relaxation oscillation problem has been in successful use in "r" measuring sets.

Even after steps were taken to stabilize measuring equipment against relaxation oscillations due to emitter-to-ground capacitances, many point-contact transistors (even some with quite low cutoff frequencies) were still unstable in the modified measuring sets. The reason for this proved to be negative resistance regions in the r_{22} or collector resistance characteristics. Such a negative resistance region is illustrated in the plot of I_c versus V_c with I_e constant, given in (a) of Fig. 4. The slope of this curve at any point is, by definition, the resistance, r_{22}, which is substantially the same as r_c, the collector resistance. Resistance r_{22} is plotted qualitatively in (b) of Fig. 4 as a function of V_c. It is seen that there is a wide region of V_c where the collector resistance is negative. If a constant-current or infinite-impedance load line is drawn on the I_c-V_c characteristic of (a) of Fig. 4, we see that for certain values of I_c, such as I_{c1}, the load

[4] H. W. Bode, "Network Analysis and Feedback Amplifier Design," D. Van Nostrand Co., New York, N. Y., p. 111; 1945.
[5] J. V. Uspensky, "Theory of Equations," McGraw-Hill Book Co., New York, N. Y., p. 304; 1948.

line intersects the curve at more than one point. This means that, if a constant-current collector bias is supplied to the transistor represented by this collector characteristic, there are three different collector voltages, V_{c1}, V_{c1}', and V_{c1}'', which are possible for the single collector current I_{c1}. It is therefore possible, as I_c is increased under the control of a constant-current bias supply, to have the collector voltage snap between V_{c1}, V_{c1}', and V_{1c}''.

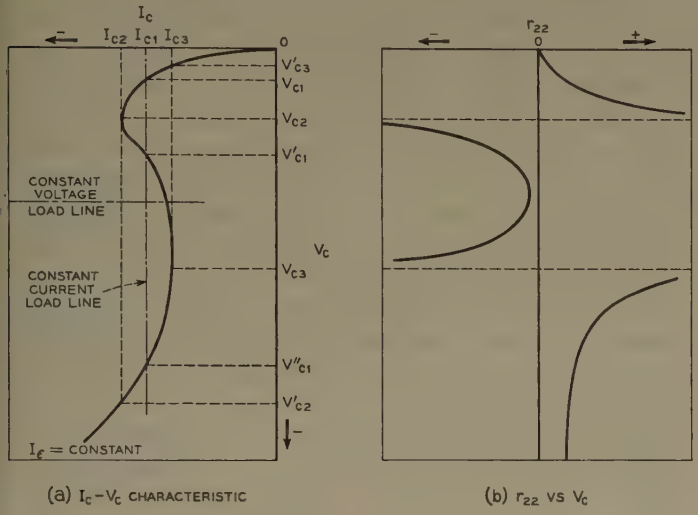


Fig. 4—Collector static characteristic showing negative resistance region. (a) $I_c - V_c$ characteristic. (b) r_{22} vs V_c.

A transistor having this type V_c-I_c characteristic is said to be open-circuit-collector unstable in the region where V_c is multivalued for a single value of I_c. Once the current increases to I_{c2}, under the control of a constant-current collector bias, the snap action referred to above is inevitable. The controlling current generator still calls for an increase in I_c, and the only way this is possible is for the collector voltage to snap from V_{c2} to V_{c2}'. Similarly, as current is decreased from a high value to a low value, the voltage must snap from V_{c3} to V_{c3}' as soon as the constant-current bias gets smaller than I_{c3}.

The current-voltage relationship, in moving from I_{c2}, V_{c2}, to I_{c2}, V_{c2}', and from I_{c3}, V_{c3}, to I_{c3}, V_{c3}', depends upon the transient response of the measuring and biasing circuits, including the transistor. As a result, high-level impulses are possible which may be sufficiently severe to damage the transistor seriously. The static V_c-I_c characteristics shown in Fig. 5 were taken on a sixty-cycle static characteristic sweeper modified to prevent relaxation oscillation due to emitter-to-ground capacitance but still using constant-current emitter and collector bias supplies. The transient impulses just referred to are very much in evidence. These transients are the same phenomena which were observed in earlier point-contact transistors and which were called "breaks" or "discontinuities" in the static characteristics. The reason they are so much more severe than those observed earlier is because the alpha cutoff frequency of the transistor, the static characteristics of which are given in Fig. 5, is in the vhf region, whereas the transistors on which similar transient phenomena or breaks had been observed earlier were an order of magnitude lower in frequency response.

Determination of True Static Characteristics

The way in which transients of the type shown in Fig. 5 can be avoided and the true static characteristics obtained, can be shown by returning to (a) of Fig. 4. A constant-voltage or zero-impedance load line has also been drawn on the curve. It is seen to intersect the I_c-V_c curve at a single point as the load line is moved up and down the voltage axis. The characteristic is therefore said to be short-circuit-collector stable. (This holds only if the emitter-to-ground impedance is sufficiently high to prevent relaxation oscillations under the conditions of a short-circuited collector as discussed above.) It would therefore be expected that, if the constant-current collector bias were replaced by a constant-voltage bias supply, the I_cV_c curve could be traced without transient breaks.

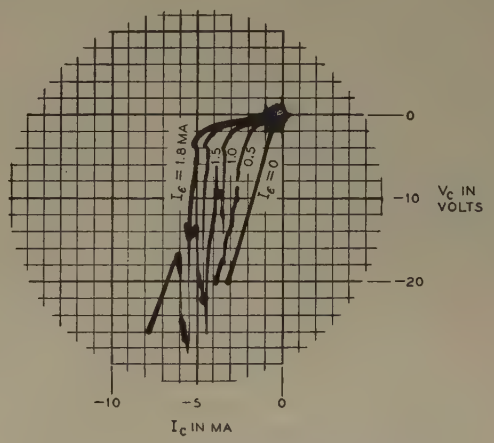


Fig. 5—Collector static characteristics of the M1832 showing negative resistance oscillations.

In order to demonstrate the effectiveness of this solution, it is illustrated in two steps. First, the circuit used in obtaining the characteristics of Fig. 5 was modified by the single addition of a 100-$\mu\mu$f capacitor bridged directly between the collector and base of the transistor. This made the collector load impedance low compared to the collector resistance of the transistor at frequencies above ten megacycles, and thereby cut the gain of the transistor in the vhf region. The static curves of Fig. 5 were then repeated, and the resulting curves are in Fig. 6, p. 1640. The extreme transients are now missing, but small breaks or discontinuities still occur in the curves. These curves serve to illustrate an important point. To have an effective constant-voltage circuit, the impedance of the collector bias supply must be very low compared to the collector resistance of the transistor being measured across the entire band of frequencies at which the transistor has forward gain. The resulting static characteristics, when this condition is met, are shown in Fig. 7. Here, the constant-current collector bias supply has been replaced by a constant-voltage supply, to assure a load impedance which is low com-

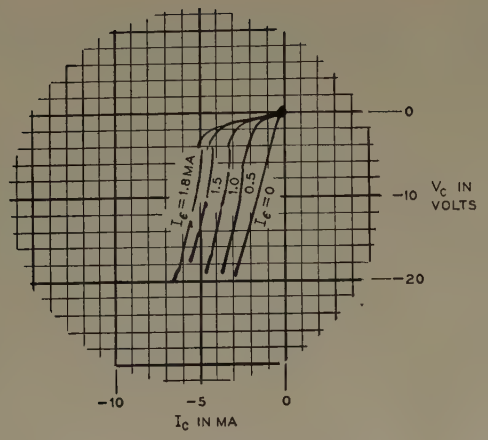


Fig. 6—Static characteristics of Fig. 5 repeated with 100 $\mu\mu$f connected between collector and base.

pared to the collector resistance from dc to the lower radio-frequency region, and a 1,000-$\mu\mu$f capacitor was bridged directly between the transistor collector and base, to assure a low collector load impedance throughout the entire radio-frequency range. The resultant static characteristics are now seen to be continuous, and they accurately trace out the negative resistance regions of the collector characteristics. The curves of Fig. 7 were actually recorded on an X-Y recorder. Comparable characteristics taken on a sixty-cycle sweeper are not presented because the unit represented by these curves was permanently damaged in an unstabilized

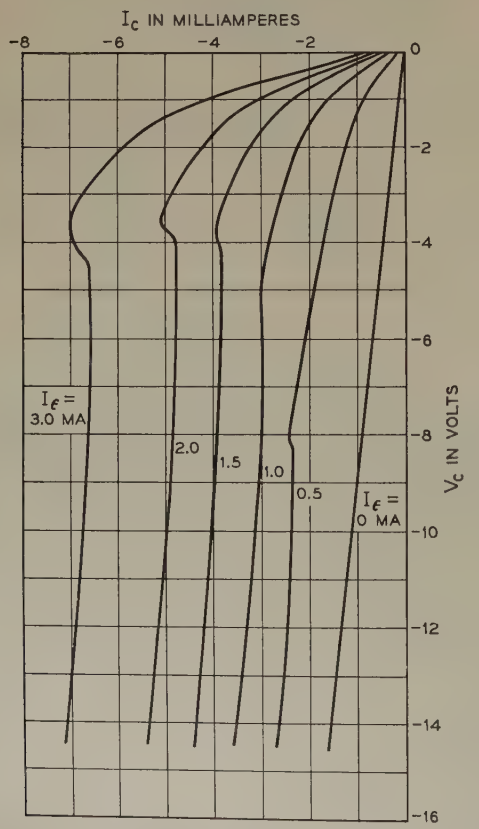


Fig. 7—Collector static characteristics of the M1832 transistor HH428 using constant-current emitter bias and constant-voltage collector bias.

circuit by transient impulses, such as those illustrated in Fig. 5, before the stabilized sixty-cycle curve tracer, using oscilloscopic presentation, was completed. However, the type of recorder used to trace the curves is immaterial once the measuring circuits have been stabilized as described above.

A schematic circuit showing one physical embodiment of the stabilization techniques described above for static-characteristic curve-tracing equipment is given in (a) of Fig. 8. Consider first the stabilization of the circuit against relaxation oscillation. Two resistances of $3K$ or greater are attached directly to the transistor emitter terminal ($1.5K$[6] is large enough to be high compared to both the emitter and base resistance of most point-contact transistors throughout their usual operating range). One of these resistances is used to feed the emitter bias current to the transistor, and the other is provided to take off emitter potential V_ϵ when needed. These resistances are used in accordance with the relaxation oscillation stabilization requirement discussed above. A variable resistance, R_ϵ, which is required to be very high compared to r_ϵ for all conditions of bias, is provided to assure a constant-current emitter bias and yet permit variations of emitter current. A resistance $R_{I\epsilon}$ is provided in series with the ground side of the emitter supply to provide a potential $V_{I\epsilon}$ for measuring the magnitude of emitter current. On the collector side, a constant-voltage supply across the entire frequency range of the transistor is assured by the 0.001-microfarad rf by-pass capacitor from collector to base[7] and by the low values of R_{vc} and R_{Ic}, which are forty ohms and ten ohms respectively. A collector-potential lead V_c is then taken directly off the collector, and a collector-current lead is taken off the high side of R_{Ic} at V_{Ic}. Emitter-current and collector-current dc meters are provided as illustrated. It is required that the collector-current meter have a sufficiently low impedance, throughout the frequency region where the 0.001-microfarad capacitor does not effectively short circuit the collector, so that the sum of its impedance and R_{vc} plus R_{Ic} is low compared to the collector resistance. To complete the circuit, suitable emitter-voltage and collector-voltage sources are supplied to the E_ϵ, E_ϵ', and E_c and E_c' terminals respectively, and suitable indicating equipment is provided to measure the voltages and currents at the V_ϵ, I_ϵ, V_c, and I_c terminals as required.

If one of the relatively slow response (response measured in seconds) commercial X-Y recorders is to be used for recording the static characteristics, suitable collector and emitter potentials can be provided by the battery circuits of (b) of Fig. 8. If sixty-cycle sweeping

[6] If the base resistance of the transistor is unusually high, these two resistances can be increased in value with no loss in effectiveness of the measuring circuit.

[7] The rf by-pass capacitor may be connected to the base in sweeping circuits only if the ac currents flowing in this capacitor are negligibly small compared to the ac currents in R_{Ic} at frequencies up to the maximum frequency required for a true oscilloscopic presentation of the collector current as a function of time. Otherwise, the capacitor should be connected from the collector to the junction of R_{vc} and R_{Ic}, i.e., at the V_{Ic} point.

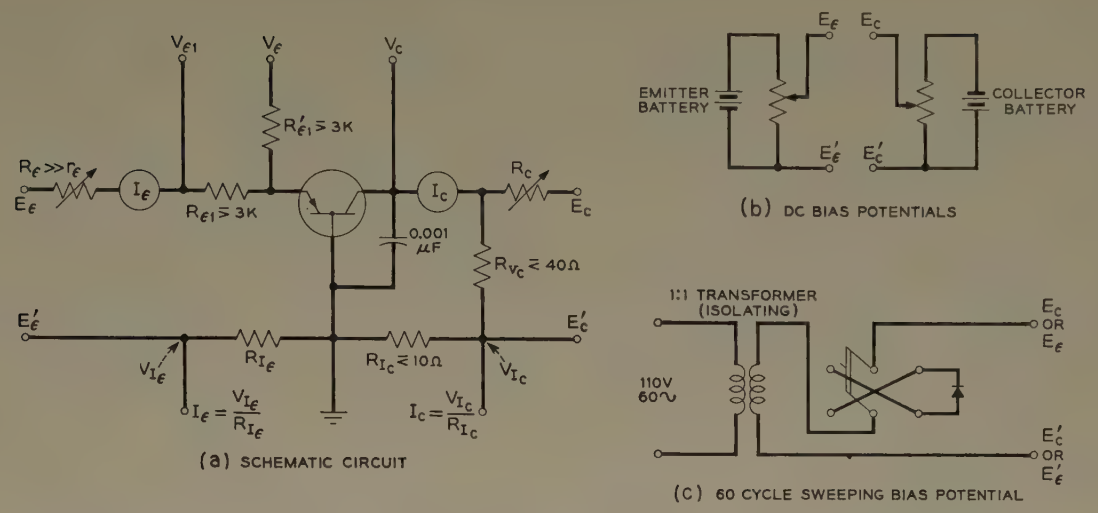


Fig. 8—Circuit for measurement of point-contact transistor static characteristics. (a) Schematic circuit. (b) 60-cycle sweeping bias potential.

bias supply and oscilloscopic presentation are desired, then the appropriate bias supply of (b) of Fig. 8 is used for the fixed bias, and the sixty-cycle bias supply of (c) of Fig. 8 is used for the independent variable bias supply. The diode of (c) of Fig. 8 is supplied to open the circuit to the unwanted polarity half-cycle of the sixty-cycle output of the 110-volt transformer.

Because of the stability problems described above, we have substituted constant-voltage collector bias for constant-current collector bias in measuring transistors having open-circuit-unstable negative resistance collector regions. We have therefore ruled out two of the open-circuit r's of Fig. 1, namely r_{11} and r_{21}, as possible parameters to be measured directly. Since it is agreed that there is nothing unique about the choice of open-circuit resistances, two short-circuit parameters can be substituted for the two open-circuit ones eliminated. The two selected are shown below:

$$h_{11} = \left.\frac{\partial V_\epsilon}{\partial I_\epsilon}\right]_{V_c=\text{constant}} = r_\epsilon + (1-\alpha)r_b$$

$$h_{21} = \left.\frac{\partial I_c}{\partial I_\epsilon}\right]_{V_c=\text{constant}} = -\frac{r_{21}}{r_{22}}. \qquad (7)$$

Since these two short-circuit parameters are directly related to the open-circuit parameters of Fig. 1, as shown in equation (7), it is possible to determine the r's of Fig. 1 and specify the transistor in terms of the equivalent circuit of Fig. 1, even though it did not prove convenient to measure two of these open-circuit resistances.

Four parameters that can be conveniently measured under stable conditions for open-circuit collector-unstable point-contact transistors are then as follows:

$$h_{11} = \left.\frac{\partial V_\epsilon}{\partial I_\epsilon}\right]_{V_c=\text{constant}}$$

$$r_{12} = \left.\frac{\partial V_\epsilon}{\partial I_c}\right]_{I_\epsilon=\text{constant}} \qquad (8)$$

$$h_{21} = \left.\frac{\partial I_c}{\partial I_\epsilon}\right]_{V_c=\text{constant}}$$

$$r_{22} = \left.\frac{\partial V_c}{\partial I_c}\right]_{I_\epsilon=\text{constant}}.$$

To get the static characteristic, the slope of which is h_{11}, V_c is set at the desired constant value using the collector battery source of (b) of Fig. 8. I_ϵ is then varied over the desired range, either by means of the battery source of (b) of Fig. 8 if an X-Y recorder is used, or by means of the sixty-cycle sweeping source of (c) of Fig. 8 if oscilloscopic presentation is to be used. The current axis of the recording instrument, with an appropriate calibration, is connected to the I_ϵ terminal, and the voltage axis (properly calibrated) is connected to V_ϵ terminal. As I_ϵ is varied, a stable curve of V_ϵ versus I_ϵ is then plotted, the slope of which at any point gives the magnitude of h_{11} at this point. Similarly for the r_{12} characteristic, I_ϵ is fixed and I_c is varied, but now under the control of a low-impedance constant-voltage source; and the recorder or oscilloscope axes are connected to the V_ϵ and I_c terminals of Fig. 8. As V_c, and consequently I_c, is varied, the V_ϵ-I_c curve is plotted, and the slope of this curve is the magnitude of r_{12}.[8] Parameters r_{22} and h_{21} are similarly obtained.

Fig. 9, page 1642, gives V_c-I_c or r_{22} characteristic of a vhf point-contact transistor having numerous negative resistance regions in its collector characteristics. These curves were taken using circuit of (a) of Fig. 8, the biasing potentials of (b) of Fig. 8, and an X-Y recorder for directly plotting the curves. It will be noted that the curves contain no discontinuities or breaks, but show I_c as a continuous function of V_c. The r_{12} static characteristic curves for the same transistor plotted with the X-Y recorder are given in Fig. 10. Here again the curves have regions of negative resistance through which the recorder plotted the V_ϵ-I_c curves continu-

[8] By shifting the axis of the recorder or oscilloscope which is connected to the I_c terminal from that terminal to the V_c terminal, the V_ϵ-V_c curve with I_ϵ constant is traced out, and the slope of this curve is the magnitude of h_{12}.

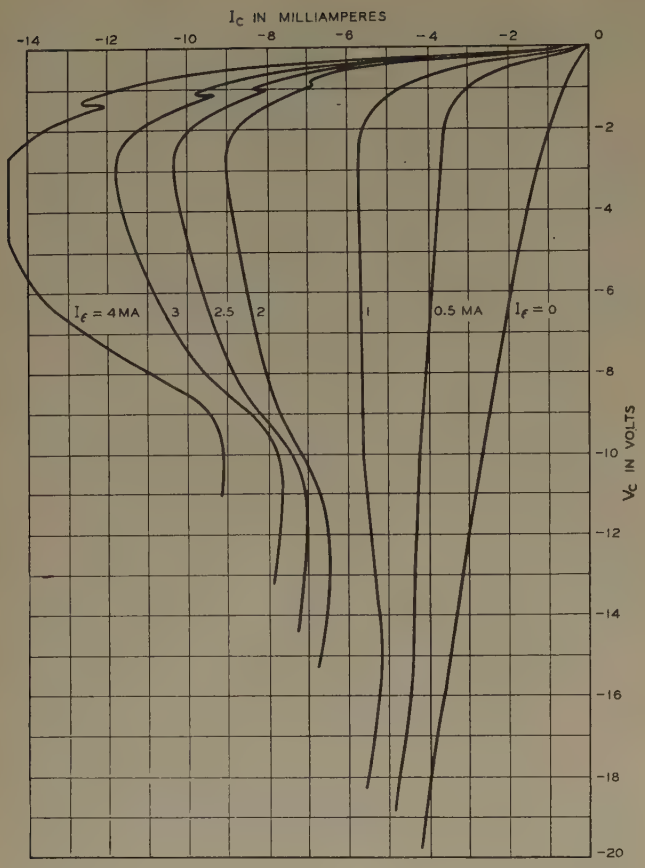


Fig. 9—Collector static characteristics of the M1832 transistor HL575 taken on the stabilized *X-Y* recorder curve tracer.

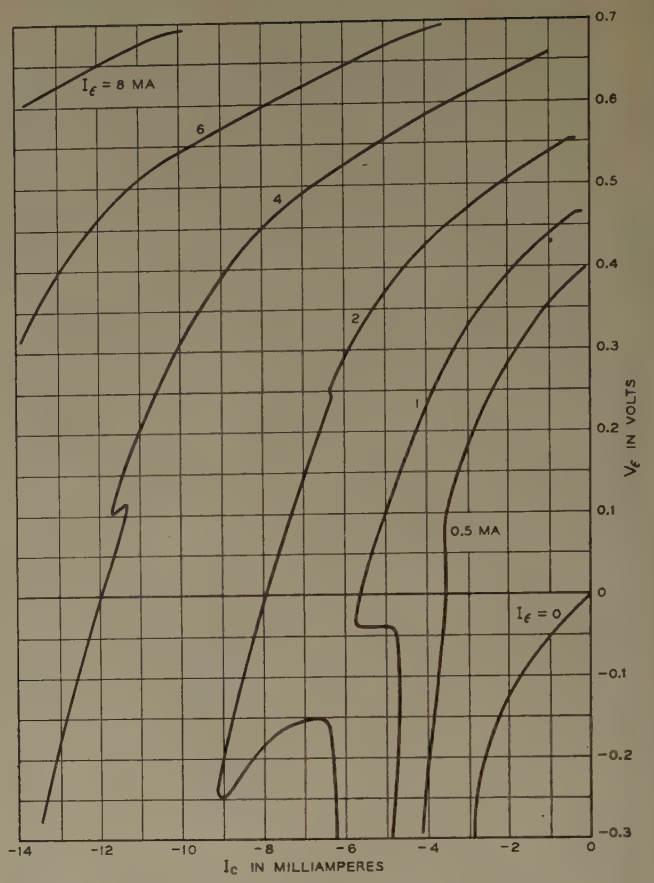


Fig. 10—Base resistance static characteristics of the M1832 transistor HL575 taken on the stabilized *X-Y* recorder curve tracer.

ously without discontinuities or breaks. The r_{22} and r_{12} curves are the more important ones and are therefore shown here for illustration. Curves for h_{11} and h_{21} may be similarly obtained, since the measuring circuit is now stable for all the conditions required by the definitions of (8). Actually, the two sets of curves given are sufficient, since all other possible parameters can be derived from these two sets of curves.

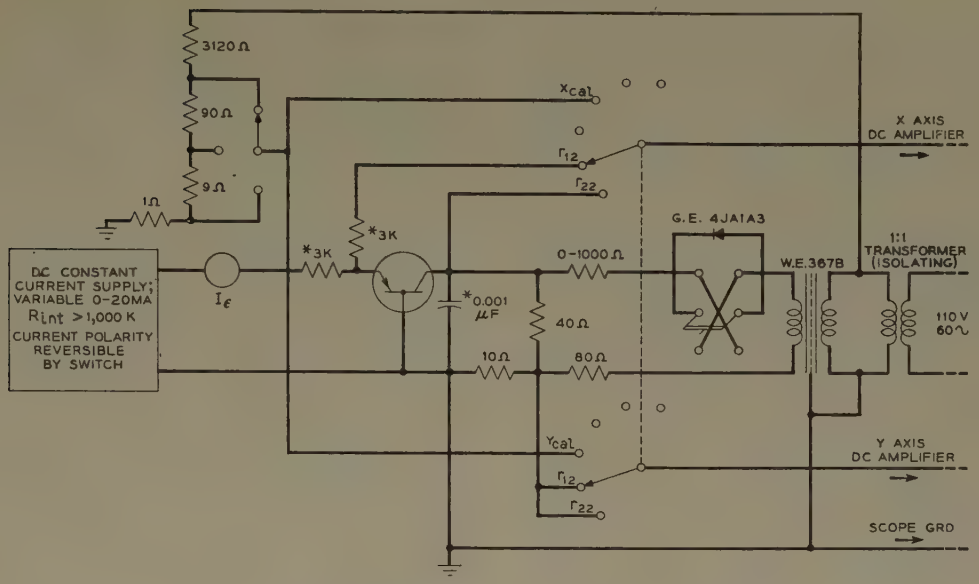
A Sixty-Cycle Sweeper

A complete sixty-cycle static characteristic sweeper, using oscilloscopic presentation which was designed according to the principles discussed previously and illustrated in Fig. 8, is shown in (a) and (b) of Fig. 11. The sweeper can be used to obtain all four characteristics defined in (8). Fig. 12 gives the r_{22} and Fig. 13 gives the r_{12} static characteristics taken with the sixty-cycle sweeper on the same transistor, the r_{22} and r_{12} characteristics of which, taken on an *X-Y* recorder, are shown in Figs. 9 and 10 respectively.

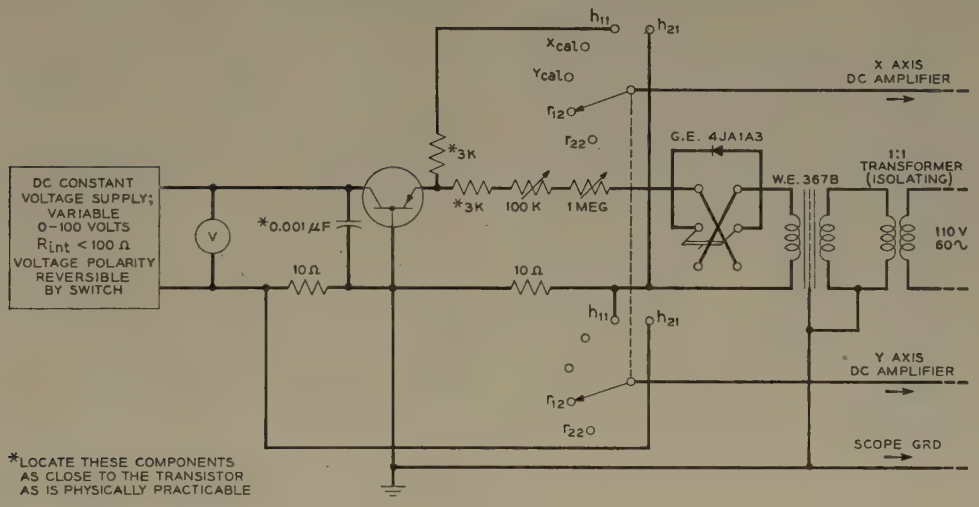
All of the circuits presented illustrate the principle of proper choice of transistor termination for stable measurements of the parameters of point-contact transistors having negative resistance regions in their collector characteristics. The important consideration is a choice of bias and signal sources having impedances for which the complete measuring circuit is stable. The actual bias supplies of Fig. 8 are inefficient, in that very high emitter batteries are necessary to assure a sufficiently high emitter bias impedance. Similarly, very high collector battery currents are needed to assure a sufficiently low collector bias impedance. Better and more convenient high-impedance emitter and low-impedance collector supplies may be obtained electronically by the use of either transistors or electron tubes combined with negative feedback. Circuits to do this have been designed, and some of them have been tried and demonstrated. In the interest of simplicity they are not presented, since they are not necessary to illustrate the principles of proper termination for stabilized measurements.

Measurement of Other Parameters

These discussions have used static-characteristic curve tracers to illustrate the principle of proper termination for stability. However, the principles are not limited to static-characteristic curve tracers but apply equally well to all parameter measurements, including direct reading "r" measuring sets, alpha sweepers, etc. Actually, as pointed out, the first use of proper stabilization against relaxation oscillation was in one of the original r measuring sets. Consider the direct reading measurement of r_{12}, for example. The dc biases would



(a) CIRCUIT FOR TRACING r_{22} AND r_{12} STATIC CHARACTERISTICS



(b) CIRCUIT FOR TRACING h_{11} AND h_{21} STATIC CHARACTERISTICS

Fig. 11—Schematic diagrams of sixty-cycle sweeper for tracing static characteristics. (a) Circuit for tracing r_{22} and r_{12} static characteristics. (b) Circuit for tracing h_{11} and h_{21} static characteristics.

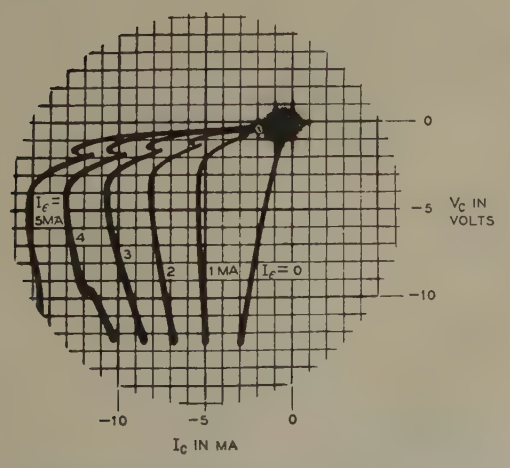


Fig. 12—Collector static characteristics of the M1832 transistor HL575 taken on the stabilized sixty-cycle sweeper.

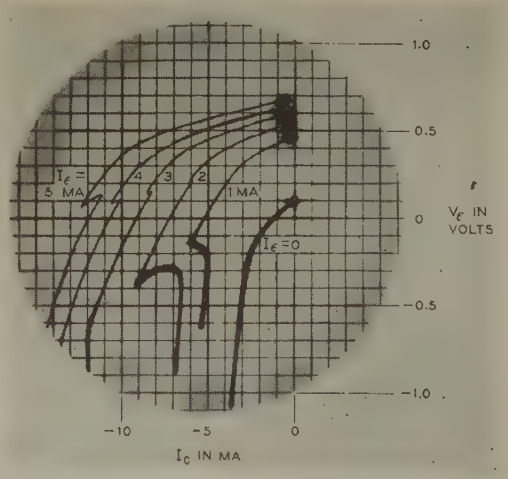


Fig. 13—Base resistance static characteristics of the M1832 transistor HL575 taken on the stabilized sixty-cycle sweeper.

be fixed by means of a constant-current emitter and constant-voltage collector supply. An ac signal generator would then be used to produce a small ac excursion of I_c about the fixed operating point. This ac signal generator would be a constant-voltage generator. The ΔI_c variation would be read as an ac voltage across a low resistance such as R_{Ic} of Fig. 8(a). The important consideration is that the impedance seen by the collector of the transistor looking into the dc bias supply, the signal generator, and the current resistor R_{Ic}, be low, not only at dc and at the signal frequency, but at all frequencies at which the transistor being measured has gain. The resultant V_e variation, ΔV_e, would then be measured as an ac voltage by a high-impedance voltmeter at a point comparable to the V_e point in Fig. 8(a). Here again it is important to assure that the impedance facing the transistor emitter is as high as possible not only at dc and at the signal frequency, but at all frequencies at which the transistor has gain. The value of r_{12} will then be given by the ratio of ΔV_e to ΔI_c.

Furthermore, although the discussion has been limited to point-contact transistors having negative resistance regions in their collector characteristics, the parameter measurements of (8) are preferable to those of Fig. 1 for high-resistance-collector point-contact transistors without negative-resistance regions and for the very high-resistance-collector junction transistors. These units will not be unstable with high-impedance collector terminations, but they will place severe requirements on a constant-current collector source because of their high collector resistance. Furthermore, although stable, they will be extremely sensitive to minute disturbances such as contact noise, etc., in the measuring circuits. This conclusion has been experimentally verified by demonstrating the superiority of the static-characteristic curve tracers discussed above for measuring such units over the older constant-collector current static characteristic tracers.

Finally, this discussion has been limited to open-circuit-emitter, short-circuit-collector terminated units. Units have been encountered that are both open-circuit and short-circuit unstable throughout certain regions of their collector characteristics. These units are usually "sports." However, they serve to illustrate an important point. A terminating impedance intermediate between open and short may be found for which such a unit is stable in the region of simultaneous open- and short-circuit instability. If this termination is used, stable measurements can be made in this region. This has been experimentally verified and further illustrates the importance of proper termination for stability in transistor measurements.

Acknowledgment

The author wishes to express his appreciation to R. C. Prim, who assisted him in the solution of the relaxation oscillation stability requirements of the circuit of (a) of Fig. 3. He also wishes to acknowledge the help of T. E. Lommasson, who built the sixty-cycle sweeper of Fig. 11 and measured the characteristics of Figs. 5, 6, 12, and 13; and of G. P. Carey, who built the curve tracer using the $X-Y$ recorder, and measured the characteristics of Fig. 7, 9, and 10.

A Mathematical Technique for the Analysis of Linear Systems*

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Summary—**The design of a linear system often takes the form of its response in the time domain to a test function. Computation of such response, particularly in the case of feedback systems, is usually complex and time consuming. Numerical methods have been applied successfully but to date, generally, they have not been related closely to the physical system being analyzed. It is shown in this paper that the z-transformation developed originally for the analysis and synthesis of sampled-data systems is applicable to numerical solutions of continuous linear systems. A model of the continuous linear system is devised in which sampling is introduced at some convenient point. The sampled time function is then reconstructed into a polygonal approximation by means of a holding operator and the output of the system is readily computed as a train of pulses giving the values of the output at sampling instants. Both analytic and arbitrary inputs can be handled by this model. The errors in the output which result by making computations on the sampled model rather than the actual system are obtainable by the application of a procedure quite similar to that of the main computation. An illustrative example using feedback is given where it is shown that the theoretical predictions concerning error produced by the model are within very close tolerances to the actual error. The system of analysis described in this paper is particularly useful to engineers because the procedures are simple, are related to the physical problem and can be applied by use of z-transform tables.**

Introduction

THE SPECIFICATION of a linear system or network often takes the form of its response in the time domain to a test function. A common test function is the unit step or ramp function although there are many cases where the input function is an arbitrary function of time. In any case, the determination of the response of the system is a relatively straightforward matter involving the solution of a set of linear differential equations. However, while straightforward, obtaining numerical values for the response as a function of time is an involved and tedious task not readily amenable to the use of desk calculators and similar computational aids. The alternative has been to apply numerical methods leading to numerical results which closely approximate the actual solution.

The application of numerical methods to the solution of linear differential equations is not new.[1] It has also been recognized that the program for obtaining a solution can be described in terms of operational procedures.[2–5] What has been lacking in previous work, however, has been an interpretation of the methods being employed in terms of the physical problem as well as a simplified and orderly technique for setting up the problem by the average engineer. By applying techniques which were developed primarily for the analysis and synthesis of sampled-data feedback control systems, this order and relationship to the physical problem can readily be established. The technique of the z-transformation,[6,7] or pulse transfer function[8] has been employed to solve problems in feedback control systems in which data are being sampled at one or more points. This established and direct approach is the one which has been applied in this paper to the numerical solution of linear systems.

The z-Transformation

In order to describe the z-transformation, reference is made to Fig. 1 where a time function $r(t)$ is sampled by means of a sampling switch S at equal intervals of time separated by a time T. The output of the switch

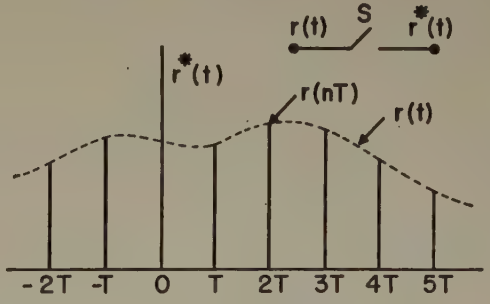

Fig. 1—Sampled time function.

$r^*(t)$ is a sequence of pulses which may be represented for mathematical purposes as a series of impulses or delta functions whose areas are equal to the amplitude of $r(t)$ at the respective sampling instants. Thus

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$$r^*(t) = \sum_{n=-\infty}^{+\infty} r(nT)\delta(t - nT). \tag{1}$$

The Laplace transform of the pulse sequence is given as

$$\mathcal{L}[r^*(t)] = \sum_{n=0}^{+\infty} r(nT)\epsilon^{-nTs}. \tag{2}$$

It is noted that the Laplace transform of such pulsed functions is in terms of ϵ^{-nTs} and it has become customary to replace this exponential by a new variable $z = \epsilon^{Ts}$ thus accounting for the name "z-transform". A very useful characteristic of these transforms is that the infinite series given in (2) may usually be expressed in closed form so that tables of z-transforms can be constructed[7,8] in much the same manner as tables of ordinary Laplace transforms.

One of the most useful properties of the z-transform is that it is possible to obtain the pulsed output of a linear system by expressing a pulse transfer function $G^*(z)$ which relates the input and output z-transforms.

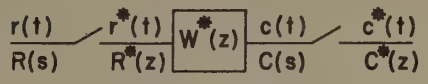

Fig. 2—Pulsed linear system.

Schematically this is shown in Fig. 2 where the output $c(t)$ is sampled synchronously with the input to give a pulse sequence $c^*(t)$. In terms of z-transforms, this relation is expressed as

$$C^*(z) = W^*(z)R^*(z). \tag{3}$$

It is readily shown that $W^*(z)$ must be the Laplace transform of the sampled output $c^*(t)$ when the input $r(t)$ is a unit impulse. The pulse transfer functions $W^*(z)$ can be expressed in closed form and tables of such transforms are available in standard works on sampled data systems.[7,8] An abridged table is given in Table I. Thus, the process of obtaining the transform of the pulsed output of a linear system subjected to a pulsed input is exactly analogous to that for continuous systems employing the ordinary Laplace transform. To obtain the output pulse sequence in the time domain, the z-transform $C^*(z)$ can be inverted by contour integration, by reference to available tables or far more simply for the purpose at hand, by a process of simple long division. This will be demonstrated later in an illustrative example.

Feedback systems can be treated by this technique by deriving relationships between the pulsed input and pulsed output for various configurations.[7] One such configuration which is of particular value to the application being discussed in this paper is shown in Fig. 3(b). The relation between the z-transform of the input and output is given by:

$$C^*(z) = \frac{RG^*(z)}{1 + HBG^*(z)} \tag{4}$$

where $RG^*(z)$ signifies the z-transform of $R(s)G(s)$ and $HBG^*(z)$ the pulse transfer function corresponding to $H(s)B(s)G(s)$. It is noted in passing that the z-transforms corresponding to $R(s)G(s)$ or $H(s)B(s)G(s)$ are not equal to the products $R^*(z)G^*(z)$ and $H^*(z)B^*(z)G^*(z)$, respectively. The whole transform $R(s)G(s)$ or $H(s)B(s)G(s)$ must be looked up in the tables. Inversion of (4) yields the sampled output time function in response to an input $r(t)$.

Application of Method to Continuous Systems

The application of the z-transform technique to the transient solution of a linear continuous system will be illustrated by applying it to feedback systems since this is where the most saving in complexity occurs. The technique used is to replace the continuous system shown in Fig. 3(a) with the sampled model of Fig. 3(b) whose response adequately approximates the actual response. The system is assumed to be low-pass so that

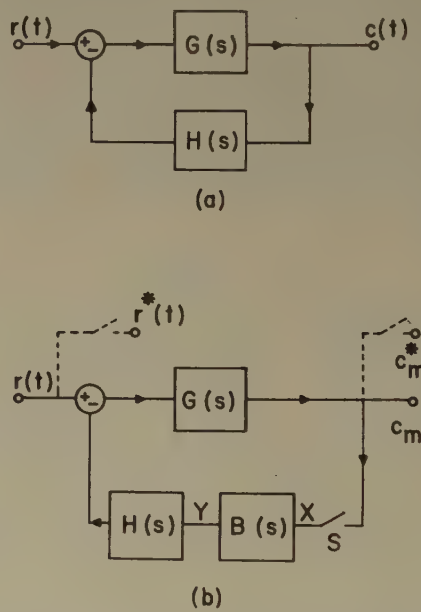

Fig. 3—(a) Continuous feedback system. (b) Sampled model for computation.

the output $c(t)$ or $c_m(t)$ contains but few high frequency components. The output is sampled by S at a frequency which is high enough to cause the error introduced by the model to be small. This requires a choice of sampling frequency at least twice the highest frequency component expected at the output.

It is assumed that the frequency at which the system's frequency response is down approximately 40 db marks the highest expected output frequency. At this frequency, only the forward gain need be considered. Therefore, as a general rule the sampling frequency should be chosen such that it is at least twice the frequency for which the forward gain is down approximately 40 db.

The output $c(t)$ in Fig. 3(b) which is fed back to the input through the switch S is a pulse sequence at the

point x and bears little resemblance to the output which is fed back in the actual system. To reconstruct more closely the actual function fed back, an element B is introduced. B can be chosen to reconstruct the original fed-back function to as high a degree of accuracy as desired. However, to prevent undue and generally unwarranted complexity a simple element is introduced which reconstructs a polygonal approximation to the actual function by converting each pulse into a triangular generating function as shown in Fig. 4. This technique,[2,3] can be represented by a block described by a transfer function $B(s)$. The requirement on $B(s)$ is that its response to a short pulse or, in the limit, an impulse of the same area, be a triangle as shown in Fig. 5(a).

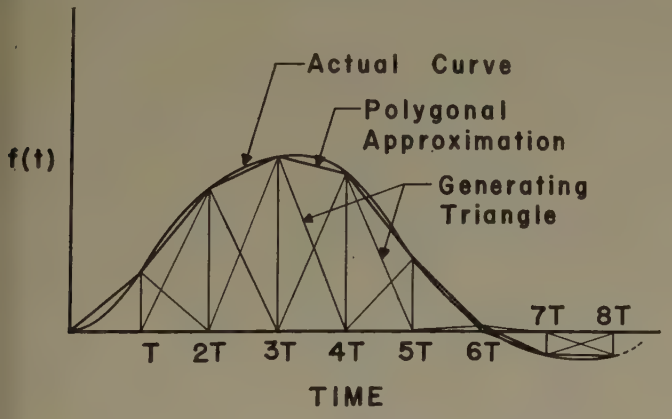

Fig. 4—Polygonal approximation of time function.

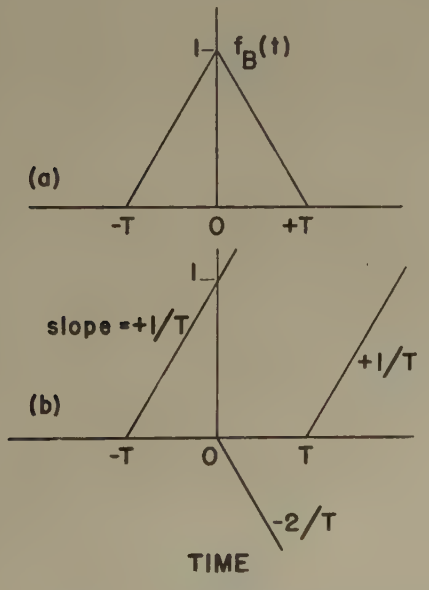

Fig. 5—Impulsive response of triangle generator.

Such a response is not physically realizable since it is initiated at a negative time but this is of no concern since the model is only mathematical in nature. The combination of the sampling switch operating synchronously with a period T and the triangle function generator B replace the direct connection which existed in the actual system. Any errors in computation resulting from use of the model are caused by the failure of this combination to faithfully reproduce the output $c(t)$ in the feedback line.

The transfer function of the triangle function generator is obtained by taking the Laplace transform of the triangle time function in Fig. 5(a). It is observed by referring to Fig. 5(b) that this impulsive response is the sum of three ramps, one advanced by the sampling time T and having a positive slope of $1/T$, the second, a ramp at the origin having a negative slope of $2/T$ and third, a ramp delayed by a sampling time T having a positive slope of $1/T$. The Laplace transform of these three components is

$$B(s) = \frac{1}{Ts^2}\epsilon^{Ts} - \frac{2}{Ts^2} + \frac{1}{Ts^2}\epsilon^{-Ts}, \tag{5}$$

which becomes

$$B(s) = \frac{\epsilon^{Ts}}{Ts^2}(1 - \epsilon^{-Ts})^2. \tag{6}$$

The Laplace transform of the loop gain of a feedback system or the cascaded gain of an open-loop system is the product of the transform of the various components. Applying the z-transformation to these components as indicated in (4) is an operation which yields the pulse transfer functions for which tables are available. For transforms which contain only powers of s and exponential terms such as those in (6) the z-transformation is quite simple as shown by reference to Table I.

Thus, if it is desired to obtain the response of a feedback system such as that of Fig. 3(a) to a test function whose Laplace transform is $R(s)$, the first step is to replace it with a model as shown in Fig. 3(b). A choice of sampling frequency $1/T$ is made using the guides previously described. The pulse transfer functions $RG^*(z)$ and $HBG^*(z)$ are then computed. These will generally be in the form of a ratio of polynomials in z^{-1}. To obtain the output of the model system at sampling instants, substitution is made in (4) and an inversion of the output z-transform $C^*(z)$ is carried out. This may be done using standard inversion theorems, but for this purpose, it is far more useful to first expand $C^*(z)$ into a power series in z by long division. $C^*(z)$ is then in the form

$$C^*(z) = c_0 + c_1z^{-1} + c_2z^{-2} + c_3z^{-3} + c_4z^{-4} + c_5z^{-5} + \cdots c_nz^{-n} + \cdots . \tag{7}$$

The magnitude of the output of the model system at sampling instants is simply the coefficient of each of the terms in the expansion at a delay corresponding to the instant in question. As an example, the output at the fifth sample time is c_5 in (7).

Since the z-transform of the output pulse sequence $C^*(z)$ is the ratio of two polynomials in z, the process of expansion of $C^*(z)$ into a power series in z is carried out by long division. This may be done either by hand or by means of a desk calculator. The process may be terminated whenever the desired number of points is ob-

tained. This procedure is clarified by means of the illustrative example which follows later. An important advantage which results from this method of inversion is that it is unnecessary to obtain the roots of the denominator of $C^*(z)$ or $C(s)$. This is generally the most time-consuming operation in the inversion procedure for higher order feedback systems. On the other hand, unless extensive tables of z-transforms are available, the roots of the internal transfer functions $G(s)$ and $H(s)$ must be found. In feedback systems, this is only a minor problem since the linear system generally consists of decoupled components having relatively simple individual transfer functions. In obtaining the response of linear networks whose transfer functions can be expressed only in terms of higher order polynomials, however, the problem of obtaining roots still remains.

Illustrative Example

To illustrate the application of the technique, the response to a step function of the system shown in Fig. 6(a) will be computed. Following the rule of selecting a

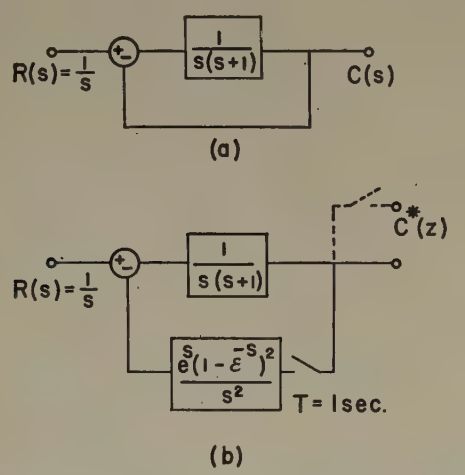

Fig. 6—System used for illustrative example.

sampling frequency at a value where the forward gain is about −40 db the convenient value of 1 sample per second is chosen. The forward gain is −30 db at this frequency. The sampling switch and triangle generator next replace the direct feedback path of the actual system as shown in the block diagram of the model in Fig. 6(b). The Laplace transform of the sampled output, $C^*(z)$ is related to the input by the z-transform relationship given in (4). The numerator of this expression, $RG^*(z)$ is the z-transform corresponding to the Laplace transform, $R(s)G(s)$:

$$RG^*(z) = \mathcal{Z}\left[\frac{1}{s^2(s+1)}\right] = \mathcal{Z}\left[\frac{1}{s^2} - \frac{1}{s(s+1)}\right]. \quad (8)$$

From Table I (at right) this becomes (for $T=1$)

$$RG^*(z) = \frac{z^{-1}}{1-z^{-1}} - \frac{z^{-1}(1-\epsilon^{-1})}{(1-z^{-1})(1-\epsilon^{-1}z^{-1})} \quad (9)$$

which upon substitution of values and simplification becomes

$$RG^*(z) = \frac{z^{-1}}{1-z^{-1}}\left[\frac{0.3679 + 0.2642z^{-1}}{(1-z^{-1})(1-0.3679z^{-1})}\right]. \quad (10)$$

Similarly, the term $HGB^*(z)$ is the z-transform corresponding to the Laplace transform $H(s)G(s)B(s)$:

$$HGB^*(z) = \mathcal{Z}\left[\frac{\epsilon^s(1-\epsilon^{-s})^2}{s^3(s+1)}\right] \quad (11)$$

which from Table I simplifies to

$$HGB^*(z) = z(1-z^{-1})^2\mathcal{Z}\left[\frac{1}{s^3} - \frac{1}{s^2(s+1)}\right] \quad (12)$$

which with further application of Table I and with the substitution of numerical values becomes

$$HGB^*(z) = \frac{0.1321 + 0.4198z^{-1} + 0.0802z^{-2}}{(1-z^{-1})(1-0.3679z^{-1})}. \quad (13)$$

Substituting the individual components from (10) and (13) into (4) and simplifying, the z-transform of the output sequence becomes

$$C^*(z) = \frac{0.3679z^{-1} + 0.2642z^{-2}}{1.1321 - 2.0802z^{-1} + 1.3962z^{-2} - 0.4481z^{-3}}. \quad (14)$$

It is recalled that this expression is the Laplace transform of the output C taken at sampling instants only. Inversion of (14) yields the values of the output at each sampling instant. This is most conveniently done by expanding (14) into a power series in z^{-1} by means of simple long division. By carrying out the long division

TABLE I

Corresponding Laplace and z-Transforms

Laplace Transform $F(s)$	z-Transform $F^*(z)$
$U(\epsilon^{Ts})V(s)$	$U^*(z)V^*(z)$
e^{-nTs}	z^{-n}
$\frac{1}{s}$	$\frac{1}{1-z^{-1}}$
$\frac{1}{s^2}$	$\frac{Tz^{-1}}{(1-z^{-1})^2}$
$\frac{1}{s^3}$	$\frac{T^2z^{-1}(1+z^{-1})}{2(1-z^{-1})^3}$
$\frac{1}{s+a}$	$\frac{1}{1-e^{-aT}z^{-1}}$
$\frac{a}{s(s+a)}$	$\frac{z^{-1}(1-e^{-aT})}{(1-z^{-1})(1-e^{-aT}z^{-1})}$
$\frac{a}{s^2+a^2}$	$\frac{z^{-1}\sin aT}{1-2z^{-1}\cos aT+z^{-2}}$
$\frac{s}{s^2+a^2}$	$\frac{1-z^{-1}\cos aT}{1-2z^{-1}\cos aT+z^{-2}}$

with the aid of a desk calculator, the z-transform of the output pulse sequence is obtained:

$$C^*(z) = 0.3250z^{-1} + 0.8305z^{-2} + 1.125z^{-3} + 1.1721z^{-4} + 1.094z^{-5} + 1.0111z^{-6} + 0.9718z^{-7} + \cdots . \quad (15)$$

The inversion of this expression into the time domain is simple since the heights of the output ordinates of the time function at each sampling instant are the coefficients of the corresponding terms in z^{-n}. These ordinates are plotted in Fig. 7. For purposes of comparison,

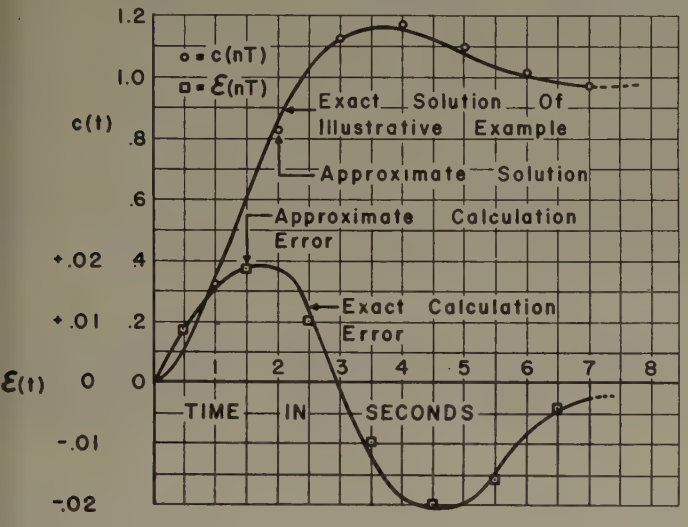

Fig. 7—Computed response and error curves for illustrative example.

the curve for the actual response obtained from the exact solution of the differential equation is also plotted. The discrepancy between the two results in no case exceeds two per cent at sampling instants where both functions are accurately known.

It is to be noted that the example worked out above is a simple one which might be solved just as well without recourse to numerical methods. However, even this simple case demonstrates that at no time was it necessary to solve the characteristic equation of the over-all system for its roots and that the operations involved are all simply handled on a calculating machine. The use of z-transform tables furthermore makes procedures used in solving the problem analogous to those normally employed in solving similar time domain problems.

Arbitrary Inputs

The previous discussions centered about those cases where the Laplace transform of the input function, $R(s)$, is known explicitly. Often, the input function is not so available but is given merely as an experimental curve or number series representing a curve which is not readily Laplace transformable. In that case, the procedures outlined previously can be applied directly by using the number or sample sequence in place of $R^*(z)$. For instance, if an arbitrary function designated by $r(t)$ were applied to the system, then the transform of the sampled function, $R^*(z)$ would be merely

$$R^*(z) = r(o) + r(T)z^{-1} + r(2T)z^{-2} + \cdots r(nT)z^{-n} + \cdots . \quad (16)$$

This cannot ordinarily be expressed in closed form but would be carried as a number series with as many terms as desired in the expressions for the output sequence transform.

For use in such cases, the computation models which more conveniently apply are shown in Fig. 8 for the open and closed circuit cases respectively. The triangle

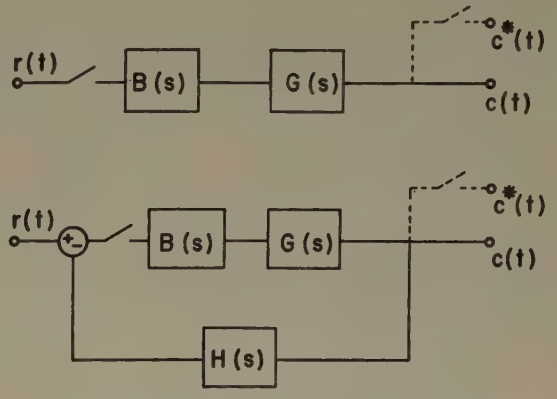

Fig. 8—Models used for arbitrary inputs.

function generator is $B(s)$ as before. For the feedback case, the location of the sampler and triangle generator shown here is not as desirable as that used with systematic inputs. The reason is that it is desirable, from the viewpoint of accuracy, to place the sampler in a position where the high frequency content of the time function is minimal. In the case of low-pass feedback systems, this generally occurs at the output of the forward transmission path. Nevertheless, the error introduced by the high frequency components of the sampler can be minimized by proper choice of the sampling frequency. For the cases shown in Fig. 8, the z-transforms of the output $c^*(t)$ are given by

$$C^*(z) = BG^*(z)R^*(z) \quad (17)$$

and

$$C^*(z) = \frac{BG^*(z)}{1 + HGB^*(z)} R^*(z) \quad (18)$$

respectively. It is noted that $R^*(z)$ appears here as a separate component and that it can be inserted independently as a power series in z^{-1} of form given in (16).

Computation Errors

One of the problems common to numerical methods is that of determining the error introduced by approximating a differential equation by a corresponding

difference equation. In cases under discussion, this resolves itself into the problem of ascertaining the error in the results introduced by a specified sampling frequency or, conversely, the determination of the sampling frequency required to obtain a specified error. Explicit solutions for sampling frequency to obtain a specified error are not, in general, known. However, an approximate error can be easily obtained for various specific system configurations, using the techniques previously described. As an example an expression for the error in the system of the illustrative example will be derived.

The definition of error can be best understood by referring to Fig. 9. Essentially, the difference between the

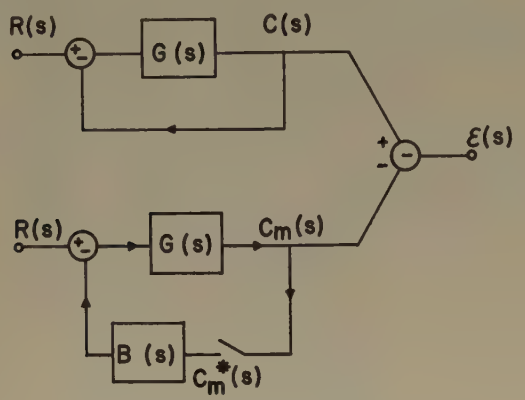

Fig. 9—System for computation of error.

output of the actual system and model system is ϵ, the computation error. The Laplace transform of this error is given by

$$\epsilon(s) = C(s) - C_m(s). \tag{19}$$

The Laplace transform of the sampled output of the model, $C_m^*(s)$ can be expanded into an infinite series[7,9]

$$C_m^*(s) = \frac{1}{T}\sum_{n=-\infty}^{+\infty} C_m(s + nj\omega_0) \approx \frac{1}{T}C_m(s). \tag{20}$$

This approximation holds for the system considered here because the contributions of other than the central term of the series, $C(s)$, are small, since the sampling frequency is chosen to be high compared to the pass band of the system.

Substituting the central term only in (19), there results for the system error

$$\epsilon(s) \approx \frac{G(s)R(s)}{1+G(s)} - \frac{R(s)G(s)}{1+\frac{1}{T}B(s)G(s)}. \tag{21}$$

Simplifying this expression and making use of the fact that $G(s)$ and $B(s)G(s)/T$ are approximately equal,

$$\epsilon(s) \approx \frac{\left[\frac{1}{T}B(s) - 1\right][G(s)]^2R(s)}{[1+G(s)]^2}. \tag{22}$$

The triangle function generator transfer function $B(s)$ is given in (6) in terms of exponential operators. If these exponentials are expanded into a power series, (22) becomes

$$\epsilon(s) \approx \frac{T^2s^2}{12}W(s)C(s) + \frac{T^4s^4}{360}W(s)C(s) + \cdots \tag{23}$$

where $W(s)$ represents the over-all transmission function

$$W(s) = \frac{G(s)}{1+G(s)}. \tag{24}$$

Consistent with the approximation made in (20) only the first term of (23) need be considered. Thus a model for obtaining the approximate computation error in a

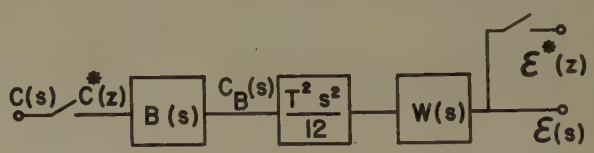

Fig. 10—Model for computing error in feedback system.

low-pass feed-back system is given in Fig. 10. The output of the triangle function generator, $C_B(s)$ is a polygonal approximation of the actual output $C(s)$ and, for purpose of error computation, differs from it only negligibly. Thus, it may be stated that

$$C(s) \approx C^*(s)B(s). \tag{25}$$

The Laplace transform of the error function then becomes

$$\epsilon(s) \approx \frac{T^2s^2}{12}B(s)W(s)C^*(s) \tag{26}$$

$$\epsilon^*(z) \approx \frac{T}{12}(z - 2 - z^{-1})C^*(z)W^*(z) \tag{27}$$

where $W^*(z)$ is the pulse transmission function of the system.

Now if the original problem being computed was the response of the system to a unit step input, $W^*(z)$ can be found with acceptable accuracy by obtaining the first back difference of the approximate output already computed. While a delay of about one-half a sampling interval is introduced this is readily eliminated by shifting forward the first difference by one-half a sampling interval. Thus,

$$W^*(z) \approx z^{1/2}\left[\frac{1}{T}(1 - z^{-1})C_m^*(z)\right]. \tag{28}$$

If this technique is applied to the illustrative problem given previously, the value of the error at sampling in-

[9] W. K. Linvill, "Sampled data control systems studied through comparison of sampling with amplitude modulation," *Trans. AIEE*, vol. 70, pp. 1779–1788; 1951.

stants is shown in Fig. 7. The error can be compared to an exact computation of error and it is noted that the results agree closely. At least insofar as the illustrative example is concerned, the general rule of sampling a low-pass system at a frequency at which the forward transmission is down approximately 40 db is a good one leading to errors in the order of a few per cent.

Conclusion

The physical interpretation of numerical techniques for obtaining solutions described by linear differential equation is a fruitful one. It leads to an understanding of the effect of the sampling procedure on the solution. In addition, the expression of numerical techniques in terms of the z-transform leads to a solution in terms of a sampled-data system analogous to the original continuous system. These sampled-data systems have been studied extensively and are in a state of development which makes their use by engineers a straightforward procedure. Thus, the numerical solution of the output of linear systems in response to either systematic or arbitrary input functions is a practical process comparable to conventional transform methods.

Traveling-Wave Amplifiers and Backward-Wave Oscillators*

MARCEL MULLER†

Summary—**This paper presents the theory of the traveling-wave tube (TWT) and magnetron amplifier, and of the corresponding backward-wave oscillators, in a unified form. Expressions are derived for amplifier gain, and oscillator starting current and tuning properties.**

Introduction

A TRAVELING-WAVE amplifier is a device in which a beam of electrons interacts with an electromagnetic wave which, by some means, has been slowed down so that its propagation is substantially synchronous with the beam. There are a number of possible realizations of such a tube: the conventional traveling-wave tube, the magnetron amplifier, the electron-wave tube, the resistance wall amplifier, etc. The most familiar of these is the conventional traveling-wave tube. Its operation has been analyzed by several largely equivalent methods.[1–4] The magnetron amplifier has received less attention until recently, and the two best-known treatments of this device[4,5] employ widely different approaches and terminologies.

It has been found that both of these types of tube, under suitable conditions of operation, can be used as oscillators of wide electronic tuning range.[6–8] Again the theory of these backward-wave oscillators is largely analogous, but no consistent exposition of it exists in the literature. This paper presents such a theory from the point of view and with the notation which is probably most familiar, that of Pierce's book.[4]

Traveling-Wave Tube and Magnetron Amplifier

The procedure commonly used to analyze the operation of the traveling-wave tube falls into two parts. These are: derivation of the field induced in the circuit by an impressed current, leading to a circuit equation; and derivation of the convection current induced in the beam by the field of the circuit. In this way one obtains two coupled differential equations which define an eigenvalue problem. The secular equation of this problem, which gives the propagation constant as a function of the frequency and of the circuit and electron beam parameters is, for the conventional TWT:

$$(\Gamma^2 - \Gamma_1^2)(j\beta_e - \Gamma)^2 = -j\beta_e\Gamma_1\Gamma^2 2C^3. \tag{1}$$

For the magnetron amplifier:

$$(\Gamma^2 - \Gamma_1^2)(j\beta_e - \Gamma)\left[(j\beta_e - \Gamma)^2 + \beta_m{}^2\right] = -j\beta_e\Gamma_1\Gamma^2\left[(j\beta_e - \Gamma) + 2j\frac{\alpha}{1+\alpha^2}\beta_m\right]H^2, \tag{2}$$

where

$$C^3 = \frac{KI_0}{4V_0}, \qquad H^2 = 2(1+\alpha^2)\phi^2C^3, \tag{3}$$

and where

Γ is the propagation constant of the system,
Γ_1 is the propagation constant of the cold circuit

$$\beta_e = \frac{\omega}{u_0}, \qquad \beta_m = \frac{\omega_c}{u_0}, \qquad \omega_c = \frac{e}{m}B,$$

* Decimal classification: R339.2×R355.912.1. Original manuscript received by the I.R.E., April 5, 1954; revised manuscript received, July 26, 1954.

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[1] J. R. Pierce, "Theory of the beam-type traveling-wave tube," Proc. I.R.E., vol. 35, p. 111; February, 1947.
[2] L. J. Chu and J. D. Jackson, "Field theory of traveling-wave tubes," Proc. I.R.E., vol. 36, p. 853; July, 1948.
[3] Bernier, *Ann. de Radioelec.*, vol. 2, p. 87; 1947.
[4] J. R. Pierce, "Traveling Wave Tubes," D. Van Nostrand Co., Inc., New York, N. Y.; 1950.
[5] O. Doehler, J. Brossart, and G. Mourier, *Ann. de Radioelec.*, vol. 3, pp. 29, 169, 328; 1948, vol. 5, p. 293; 1950.
[6] R. Kompfner and N. T. Williams, "Backward-wave tubes," Proc. I.R.E., vol. 41, p. 1602; November, 1953.
[7] P. Guenard, O. Doehler, and R. Warnecke, *Comptes Rendus*, vol. 235, p. 32; 1952.
[8] R. Warnecke, *L'Onde Electrique*, no. 325; April, 1954.

ω is the frequency,
u_0 is the beam velocity,
B is the magnetic induction,
e and m are the electron charge and mass,
I_0 and V_0 are the dc beam current and voltage,
K is the wave impedance of the circuit, and
ϕ and α are dimensionless parameters describing the transverse variation of the high-frequency field.

We note that the equation for the magnetron amplifiers is of the fifth degree, while that for the conventional TWT is of the fourth. This arises from the additional degree of freedom (transverse electron motion) allowed in the magnetron amplifier.

Now in a lossless circuit we may substitute $\Gamma_1=j\beta$, and we estimate the roots by writing $\Gamma=jk+\varphi$, where φ is a small correction (incremental propagation constant). Making these substitutions and neglecting certain small terms, we obtain

Magnetron amplifier:

$$[\beta^2-k^2+2jk\varphi][j(\beta_e-k)-\varphi][\beta_m{}^2-(\beta_e-k)^2-2j(\beta_e-k)\varphi]$$
$$=-j\beta\beta_e k^2\left\{(\beta_e-k)+\frac{2\alpha}{1+\alpha^2}\,\beta_m\right\}H^2 \qquad (4)$$

TWT:

$$[\beta^2-k^2+2jk\varphi][j(\beta_e-k)-\varphi]^2=-\beta_e\beta k^2 2C^3. \qquad (5)$$

Since the right-hand members of (4) and (5) are small, the equations will be satisfied if one or more of the factors on the left-hand side are small. This means that the roots of equation (4) will occur near $k=\pm\beta$, $k=\beta_e$, $k=\beta_e\pm\beta_m$, the roots of equation (5) near $k=\pm\beta$, $k=\beta_e$ (double root). This behavior of the approximate propagation constants is represented in Figs. 1a–1c in terms of the phase velocities of the corresponding waves; that is to say, Figs. 1a–1c show, for a fixed circuit phase velocity, the behavior of the phase velocities of the induced waves as the electron speed is varied. In order to read off the velocities of the induced waves, a vertical line is drawn at the desired value of u_0, the electron speed, as indicated in Fig. 1a.

Before proceeding to solve equations (4) and (5) for the incremental propagation constants, it is instructive to consider the nature of the solutions more closely. It can be seen by inspection of equation (4) that if only one of the factors on the left is small, φ will be purely imaginary. The possibility of complex φ is confined to regions where two or more of the factors are small. One can see similarly in equation (5), as is indeed well known, that the possibility of complex φ arises only when $\beta\cong\beta_e$ or $\beta>\beta_e$. This has been represented in Fig. 1 by emphasizing the regions of possibly complex φ.

One can now obtain the literal values of φ by assigning the appropriate values of k. It is more instructive, however, to employ a graphical representation of the analytical procedure. To obtain such a representation, one can solve both equations of the system, the ballistic

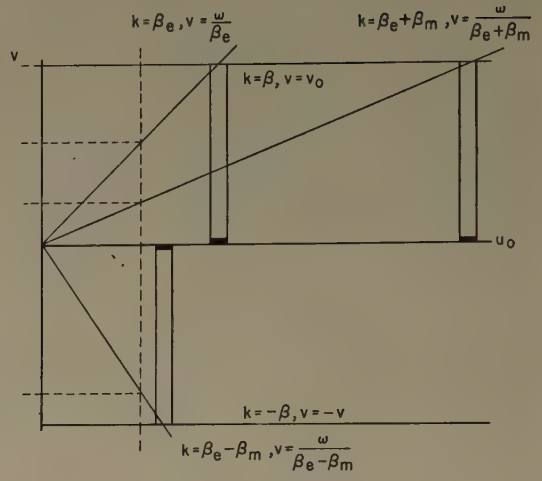

Fig. 1a—$\omega<\omega_c=eB/m$.

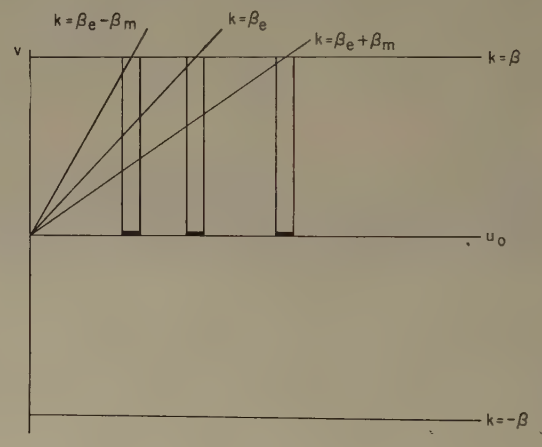

Fig. 1b—$\omega>\omega_c$.

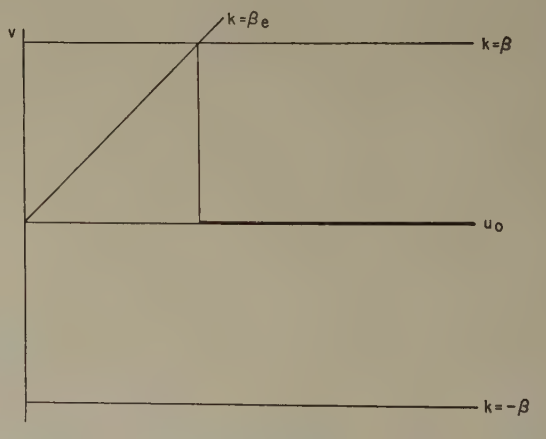

Fig. 1c—$\omega_c=0$.

and circuit equations for ρ/V, where ρ is the rf charge density and V the rf voltage. This quantity, analogous to an admittance, is then plotted as a function of propagation constant. In this way two curves are obtained the (real or imaginary) intersections of which determine the propagation constants for the system. This will be done first for the conventional traveling-wave tube

$$\frac{\rho}{V}=\frac{j(\Gamma^2-\Gamma_1{}^2)}{\omega\Gamma_1 K} \qquad \text{(circuit equation)} \qquad (6)$$

$$\frac{\rho}{V} = \frac{I_0\Gamma^2\beta_e}{2\omega V_0(j\beta_e - \Gamma)^2} \quad \text{(ballistic equation).} \quad (7)$$

If we normalize by writing $\Gamma = j\gamma$, $\Gamma_1 = j\beta$, $\delta = \gamma/\beta_e$, then

$$\frac{\rho}{V} = \frac{\left(\frac{\beta}{\beta_e}\right)^2 - \delta^2}{\frac{\omega\beta}{\beta_e{}^2}K} \quad \text{(circuit equation)} \quad (8)$$

$$\frac{\rho}{V} = \frac{I_0\beta_e\delta^2}{2\omega V_0(1-\delta^2)} \quad \text{(ballistic equation).} \quad (9)$$

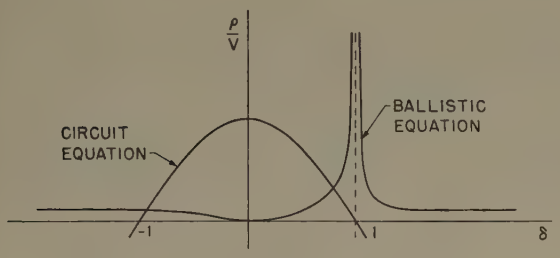

Fig. 2a—The TWT at synchronism.

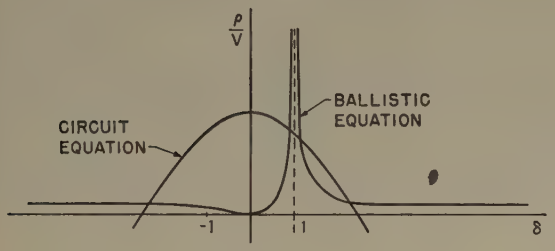

Fig. 2b—The TWT with a slow beam.

Equations (8) and (9) have been plotted in Figs. 2a and 2b. The corresponding equations for the magnetron amplifier are:

$$\frac{\rho}{V} = \frac{\left(\frac{\beta}{\beta_e}\right)^2 - \delta^2}{\omega\frac{\beta}{\beta_e{}^2}\phi K} \quad \text{(circuit equation)} \quad (10)$$

and, with $m = \beta_m/\beta_e$,

$$\frac{\rho}{V} = -\frac{nI_0\phi}{u_0{}^3}\cdot\frac{\delta^2[(1+\alpha^2)(1-\delta)+2\alpha m]}{(1-\delta)[m^2-(1-\delta)^2]} \quad \text{(ballistic equation).} \quad (11)$$

Equations (10) and (11) have been plotted in Figs. 3a–3d. By observing the changes of the figures as the magnetic field is reduced, it is easily seen how the behavior of the magnetron amplifier and of the conventional traveling-wave tube are related. Of particular interest is the fact that the electronic admittance of the magnetron amplifier, equation (11), has a simple pole at $\delta = 1$. It is this pole that gives rise to the roots of the secular equation leading to growing waves when the electron stream is near synchronism, as can be seen from Figs. 3a, 3b, and 3c. Since the pole is simple, it gives rise to

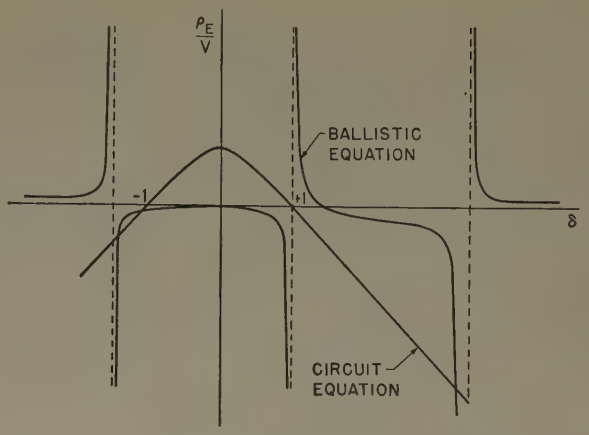

Fig. 3a—$m > 2$, at synchronism.

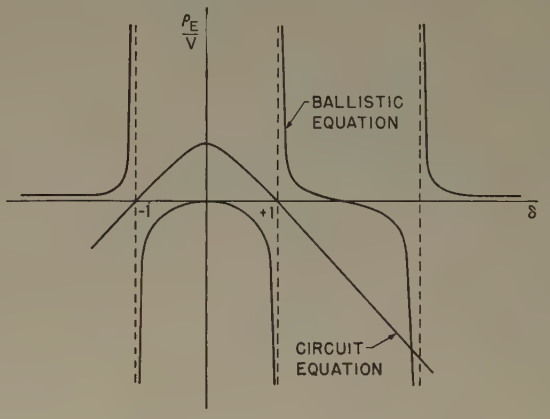

Fig. 3b—$m = 2$, at synchronism.

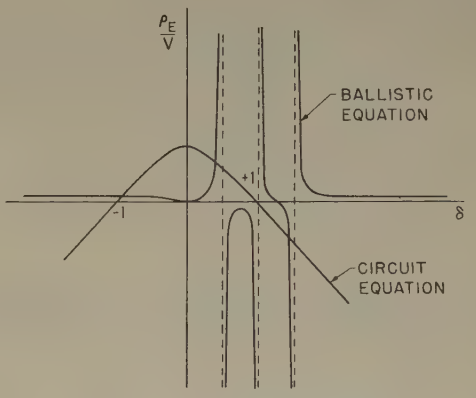

Fig. 3c—$m < 1$, at synchronism.

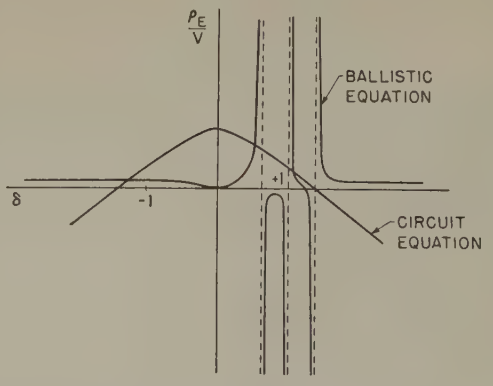

Fig. 3d—$m < 1$, off synchronism.

just two roots of the secular equation close together. As the magnetic field is reduced to zero, this simple pole vanishes and the other two simple poles at $\delta = 1 \pm m$ coalesce to form the double pole of the conventional traveling-wave tube, as in Figs. 2a and 2b. (See also Appendix II.) When a conventional traveling-wave tube is operated near synchronism (Fig. 2a), this double pole gives rise to three roots close together. These correspond to the familiar three forward waves. In order to determine the incremental propagation constants for these three waves by the approximation method used here, a cubic equation must be solved, and one might expect the roots to be proportional to $I_0^{1/3}$; in the case of the magnetron amplifier the procedure evidently leads to a quadratic equation and the incremental propagation constants will be proportional to $I_0^{1/2}$. From the figures, it is clear that growing waves can also be obtained with the beam not at synchonism. One such case is shown in Fig. 3d.

We can now consider a special situation that arises when $\beta_m = 2\beta_e$, by referring to Fig. 3b. Evidently in this case there is only one pure imaginary propagation constant, occurring at $\gamma \cong \beta_e + \beta_m$, and there can be both forward and backward amplifying waves when the beam is near synchronism. Thus it is very likely that this situation will lead to self-oscillation of the amplifier at a frequency $\omega = \frac{1}{2}\omega_c$, if this frequency is within the passband of the circuit. To determine whether this surmise is correct, it would again be necessary to deal with the boundary conditions.

In order to put the equations giving the incremental propagation constants into a form usable for the computation of gain, Pierce defines a set of normalized parameters to describe velocity separation and circuit coupling. We will proceed in exactly analogous fashion to deal with equation (4). Letting

$$\frac{\beta}{\beta_m}\frac{\alpha}{1+\alpha^2}H^2 \equiv D^2 \quad \text{and} \quad \varphi = \beta_e D\delta, \tag{12}$$

we obtain an equation for the normalized propagation constant δ, with the beam near synchronism

$$\delta(\delta - jb) = 1 \tag{13}$$

where

$$-b = \frac{1}{D}\left(1 - \frac{\beta}{\beta_e}\right), \tag{14}$$

so that b describes the departure from synchronism.

The maximum gain occurs at $b = 0$, that is, at exact synchronism, where $\delta = \pm 1$. If $|b| > 2$, δ has no real part and there is no growing wave. Similar results are obtained for the other situations in which complex roots can occur. They are of relatively little practical interest since, in practice, magnetron amplifiers are always used with the electron beam at synchronism.

We now complete the discussion of this case by inserting the boundary conditions in accordance with Appendix I. At

$$Z = 0, \quad V_1 + V_2 = V, \quad \text{and} \quad \rho_1 + \rho_2 = \rho = 0;$$

that is to say, the beam enters unmodulated. Thus

$$\frac{V_1}{\delta_1} + \frac{V_2}{\delta_2} = 0 \tag{15}$$

$$V_1 = \frac{V}{1 - \dfrac{\delta_2}{\delta_1}} \tag{16}$$

and, since at exact synchronism $\delta_2/\delta_1 = -1$,

$$V_1 = \frac{V}{2}. \tag{17}$$

In this way we have obtained the two features that most characteristically distinguish the magnetron amplifier from the conventional traveling-wave tube: first, the fact that the gain is proportional to $I_0^{1/2}$ rather than $I_0^{1/3}$; second, the value of the initial loss parameter which is 1/2, or 6.0 db, rather than 1/3 or 9.5 db.

The Backward-Wave Oscillator

The operation of the backward-wave oscillator is based on the existence of disturbances that propagate energy in a direction opposite that of their phase velocity. It is well known from the Brillouin diagram[9] that any periodically loaded structure exhibits such modes of propagation.

It is shown by Pierce that a circuit operating in such a mode may be described by a circuit equation

$$V = +\frac{\Gamma\Gamma_1 Ki}{\Gamma^2 - \Gamma_1^2}, \tag{18}$$

all other elements of the theory remaining unchanged. This leads to a new secular equation in the case of the conventional traveling-wave tube:

$$1 = -\frac{jKI_0\beta_e\Gamma^2\Gamma_1}{2V_0(\Gamma_1^2 - \Gamma^2)(j\beta_e - \Gamma)^2} \tag{19}$$

and in the case of the magnetron amplifier:

$$[\beta^2 - k^2 + 2jk\varphi][j(\beta_e - k) - \varphi] \cdot [\beta^2 - (\beta_e - k)^2 - 2j(\beta_e - k)\varphi] = +j\beta\beta_e k^2\left\{\beta_e - k + \frac{2\alpha}{1+\alpha^2}\beta_m\right\}H^2. \tag{20}$$

The requirement that the field vanish at a point L from the beginning of the circuit imposes the conditions[10]

$$(\beta - \beta_e)L = bCN = bC\beta_e L \cong (2n+1)\pi$$

and appropriate values of $CN = C\beta_e L$.

The first of these conditions determines the tuning properties of the oscillator. It can be seen that if β were constant, the frequency would be proportional to the

[9] See, e.g., J. C. Slater, "Microwave Electronics," p. 174, D. Van Nostrand Co., Inc., New York, N. Y.; 1950.

[10] H. Heffner, "Analysis of the Backward-Wave Traveling-Wave Tube," Tech. Rept. No. 48, Contract N6 Onr-251, Elec. Res. Lab., Stanford Univ., Stanford, Calif.; 1952.

square root of the voltage (in fact, of course, a backward space harmonic can never have a constant β; this means that the tuning rate will be smaller). The second condition specifies the smallest gain parameter C for which oscillation can take place, and thus the starting current, since $C^3 = KI_0/4V_0$.

The theory of the magnetron-type backward-wave oscillator is based, in an entirely analogous fashion, on the secular equation (20).

Comparison with equation (2) shows how the nature of the roots is changed in this case. Away from the critical points, the propagation constants are still purely imaginary. Near the critical points, the values of φ are changed; in fact, wherever (2) yielded a complex value of φ, (20) makes φ pure imaginary and vice versa. This can be seen in another way by referring to Figs. 3a–3d again. In deriving (20), the ballistic equation has not been changed; thus the ballistic curves in these figures remain the same. The circuit equation, however, has been changed by reversing the sign of the circuit admittance. This means that in the representation of the roots of (20), the circuit parabola must be mirrored in the x-axis. Inspection of the figures shows immediately that an imaginary intersection will now give rise to two real intersections, etc.

One consequence of this situation is that now the wave traveling near synchronism with the beam is *no longer amplified.* The incremental propagation constants in this case are given by the analog of equation (13); thus

$$\delta(\delta - jb) = -1. \tag{21}$$

Nonetheless, *oscillation can still take place;* thus

$$V(L) = V_1(L) + V_2(L) = V_1(0)e^{-\Gamma_1 L} + V_2(0)e^{-\Gamma_2 L}. \tag{22}$$

From (16)

$$V(L) = V(0)\left[\frac{e^{-\Gamma_1 L}}{1 - \frac{\delta_2}{\delta_1}} + \frac{e^{-\Gamma_2 L}}{1 - \frac{\delta_1}{\delta_2}}\right]$$

$$= V(0)\left[\frac{\delta_1 e^{-\Gamma_1 L} - \delta_2 e^{-\Gamma_2 L}}{\delta_1 - \delta_2}\right]. \tag{23}$$

Oscillation is possible when the bracket vanishes, or

$$\delta_1 e^{-\beta_e C \delta_1 L} = \delta_2 e^{-\beta_e C \delta_2 L}, \tag{24}$$

where δ_1, δ_2 are the roots of (21), so

$$\delta_{1,2} = \frac{j}{2}\left\{b \pm \sqrt{b^2 + 4}\right\} \tag{25}$$

so that

$$\frac{\delta_1}{\delta_2} = \frac{b - \sqrt{b^2 + 4}}{b + \sqrt{b^2 + 4}}, \tag{26}$$

but from (24)

$$\frac{\delta_1}{\delta_2} = e^{-\beta_e C L(\delta_2 - \delta_1)} = e^{-j\beta_e C L\sqrt{b^2+4}}. \tag{27}$$

Equation (27) is satisfied only if $b = 0$ and

$$2\beta_e CL = (2n + 1)\pi \quad n = 0, 1, 2, \cdots, \tag{28}$$

or equivalently $CN = (2n+1)/4$.

Physically, this oscillation criterion means that two unattenuated waves are propagated down the same structure at the same frequency but at slightly different propagation velocities. If the waves start in phase at the gun end of the tube, they will, an appropriate length of tube further on, be 180° out of phase and cancel exactly. In such a situation there will be a finite circuit field at the gun end of the tube and zero field at the collector end. This means that in a reverse direction the tube has infinite gain, and thus is an oscillator. Unlike a conventional traveling-wave tube, a magnetron amplifier oscillates at exact synchronism of wave and electron streams. Also in this case, it is very clearly shown that the backward oscillation and backward gain do not depend in any way on a growing wave, because in this case there is no growing wave present, but do depend on an interference phenomenon. An analogy in the ordinary traveling-wave tube is pointed out by Heffner[10] for the case of very large space charge densities, when the gain of the growing wave may be reduced to unity. The starting current of the oscillator can be computed from (28), which contains the gain parameter and thus the beam current.

The tuning properties of the tube are given by the requirement $b = 0$, i.e., $\beta = \beta_e$ or

$$\omega = \beta u_0. \tag{29}$$

If β is constant, this means that the frequency is proportional to the beam velocity. (The effect of dispersion is to make the frequency increase somewhat more slowly than the beam velocity.)

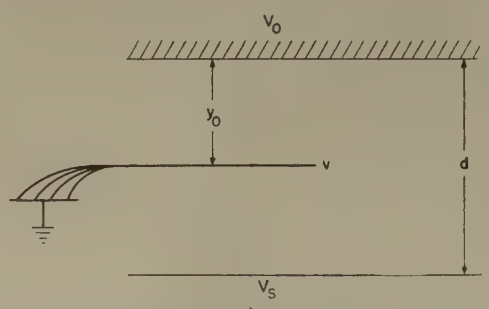

Fig. 4—Cathode geometry of magnetron amplifier.

It was pointed out above, in connection with the conventional type of backward-wave oscillator, that the frequency, aside from dispersion, is proportional to the square root of the beam voltage. In the case of the magnetron-type of backward-wave oscillator, one might take advantage of the relation

$$u_0 = -\frac{E}{B}$$

to obtain a more favorable tuning characteristic. Consider the structure of Fig. 4. The potential at the beam is V, given by

$$V = V_0 - \frac{y_0}{d}(V_0 - V_s). \tag{30}$$

From energy conservation

$$u_0 = \sqrt{\frac{2e}{m}}\, V. \tag{31}$$

Combining (29), (30), and (31)

$$\omega = \beta\sqrt{\frac{2e}{m}}\left[V_0 - \frac{y_0}{d}(V_0 - V_s)\right]^{1/2}. \tag{32}$$

Also

$$u_0 = -\frac{E}{B} = \frac{V_0 - V_s}{Bd}, \tag{33}$$

so

$$\omega = \beta\frac{V_0 - V_s}{Bd}. \tag{34}$$

Solving (34) for $(V_0 - V_s)$ and inserting in (32),

$$\omega = \beta\sqrt{\frac{2e}{m}}\left[V_0 - \frac{y_0 B}{\beta}\,\omega\right]^{1/2}, \tag{35}$$

and solving for V_0 and V_s gives

$$V_0 = \frac{v_0 B}{\beta}\,\omega + \frac{m}{2e\beta^2}\,\omega^2 \tag{36}$$

$$V_s = -(d - y_0)\frac{B}{\beta}\,\omega + \frac{m}{2e\beta^2}\,\omega^2. \tag{37}$$

On the basis of (36) and (37) it is possible to choose a variation of y_0 over the tuning range so as to maximize the frequency variation for a given change in electrode voltage. Some special cases will be discussed:

A) $y_0 = 0$

$$V_0 = \frac{m\omega^2}{2e\beta^2},$$

so $\omega \propto V_0^{1/2}$ if $\beta = \text{const}$

$$V_s = \frac{m\omega^2}{2e\beta^2} - \frac{Bd\omega}{\beta}.$$

B) $y_0 = a/\beta$ where a is some constant; in general $a \leqq 1$ to maintain the circuit impedance high.

$$V_0 = \frac{1}{\beta^2}\left[aB\omega + \frac{m}{2e}\,\omega^2\right].$$

The two terms in this equation are of comparable magnitude; thus in this case the tuning rate is intermediate between $\omega \propto V_0^{1/2}$ and $\omega \propto V$.

C) If we let the beam approach the circuit at the higher frequencies so that the efficiency decreases (which might be made up by the increased power input),

$$y_0 = \frac{a}{\beta}\,\frac{\omega}{\omega_1},$$

then

$$V_0 = \frac{1}{\beta^2}\left[a\omega_1 B + \frac{m\omega^2}{2e}\right],$$

which can give quite a steep tuning rate, and

$$V_s = \frac{1}{\beta^2}\left[a\omega_1 B + \frac{m\omega^2}{2e}\right] - \frac{dB}{\beta}\,\omega.$$

These examples serve to point out the possibilities that might be realized.

Acknowledgement

Thanks are due to Drs. R. L. Jepsen and W. L. Beaver for many helpful discussions, and to Prof. M. Chodorow for reading the manuscript.

Appendix I

One can consider the boundary conditions for the magnetron amplifier relatively easily for the case in which the output (collector) end of the tube is matched so that no energy is reflected. This means that the amplitude of any backward waves is zero.

The beam enters initially unmodulated: $\rho(0) = 0$, $\dot{z}(0) = 0$. The waves will be numbered as follows:

$$\Gamma \cong j\beta_c : 1, 2, \qquad \Gamma \cong j(\beta_e + \beta_m) : 3, \qquad \Gamma \cong j(\beta_e - \beta_m) : 4$$

With the appropriate approximations[11]

$$\rho_1 = \frac{j\alpha\beta_e n I_0}{u_0^3 D\beta_m\delta_1}\,V_1$$

similarly ρ_2

$$\rho_3 = \frac{j(\beta_e + \beta_m)^2\eta I_0(1 - \alpha^2)}{2u_0^3\beta_m\varphi}\,V_3.$$

Now from (4) with $k = \beta_e + \beta_m$

$$\varphi = \frac{j\beta_e\beta(\beta_e + \beta_m)^2\,\dfrac{(1-\alpha)^2}{1+\alpha^2}\,H^2}{2[\beta^2 - (\beta_e + \beta_m)^2]\beta_m},$$

and using $\beta_e \cong \beta$

$$\rho_3 = -\frac{(1 + \alpha^2)\beta_m(2\beta_e + \beta_m)\eta I_0}{u_0^3\beta_e\beta(1 - \alpha)H^2}\,V_3.$$

Also

$$\dot{z} = -\frac{\eta\Gamma\phi V\left[(j\beta_e - \Gamma) + j\alpha\beta_m\right]}{u_0[(j\beta_e - \Gamma) + \beta_m^2]}.$$

So

$$\dot{z}_1 = +\frac{\eta\beta_e\phi V_1}{u_0\beta_m}.$$

Similarly $\dot{z}_2$

[11] J. R. Pierce, *op. cit.*, p. 113.

$$\dot{z}_3 = \frac{j\eta(\beta_e+\beta_m)\phi(1-\alpha)}{2u_0\varphi} V_3$$

$$= -\frac{(1+\alpha^2)\eta\phi\beta_m{}^2(2\beta_e+\beta_m)}{u_0\beta_e\beta(\beta_e+\beta_m)(1-\alpha)H^2} V_3.$$

Substituting these into the boundary conditions where $\beta_m > \beta_e$ (three forward waves),

$$V_1 + V_2 + V_3 = V$$

$$\frac{V_1}{\delta_2} + \frac{V_2}{\delta_2} + \frac{j(1+\alpha^2)\beta_m{}^2(2\beta_e+\beta_m)D}{\alpha(1-\alpha)\beta_e{}^3H^2}$$

$$= -\frac{ju_0{}^3D\beta_m}{\alpha\beta_e\eta I_0}\rho = 0$$

$$V_1 + V_2 - \frac{(1+\alpha^2)\beta_m{}^3(2\beta_e+\beta_m)}{(1-\alpha)\beta_e{}^3(\beta_e+\beta_m)H^2}$$

$$= \frac{u_0\beta_m}{\eta\beta_e\phi}\dot{z} = 0.$$

Substracting the third of these equations from the first

$$V_3\left[1 + \frac{(1+\alpha^2)\beta_m{}^3(2\beta_m+\beta_m)}{(1-\alpha)\beta_e{}^2(\beta_e+\beta_m)H^2}\right] = V.$$

The bracket is a very large number; hence, $V_3 \ll V$ and this wave is not excited to any appreciable extent.

The case of small magnetic field ($\beta_e > \beta_m$) requires a slightly more involved proof; there are now four forward waves; one derives just as before

$$V_4 = \frac{u_0{}^3\beta_e{}^2(1+\alpha)H^2}{\eta I_0(1+\alpha^2)\beta_m(2\beta_e-\beta_m)}\rho_4$$

and

$$\dot{z}_4 = \frac{j\eta(\beta_e-\beta_m)\phi(1+\alpha)}{2u_0\varphi} V_4.$$

Using the appropriate value of φ

$$\dot{z}_4 = -\frac{(1+\alpha^2)\eta\phi\beta_m{}^2(2\beta_e-\beta_m)}{u_0\beta_e{}^2(\beta_e-\beta_m)(1+\alpha)H^2} V_4,$$

so

$$-\frac{j(1+\alpha^2)\beta_m{}^2(2\beta_e-\beta_m)D}{\alpha(1+\alpha)\beta_e{}^3H^2} V_4 = -\frac{ju_0{}^3\beta_m D}{\alpha\beta_e\eta I_0}\rho_4$$

$$-\frac{(1+\alpha^2)\beta_m{}^3(2\beta_e-\beta_m)}{(1+\alpha)\beta_e{}^3(\beta_e-\beta_m)H^2} V_4 = \frac{u_0\beta_m}{\beta_e\eta\phi}\dot{z}_4.$$

Substituting into the boundary conditions

$$V_1 + V_2 + V_3 + V_4 = V$$

$$\frac{V_1}{\delta_1} + \frac{V_2}{\delta_2} + \frac{j(1+\alpha^2)\beta_m{}^2(2\beta_e+\beta_m)D}{\alpha(1-\alpha)\beta_e{}^3H^2} V_3$$

$$-\frac{j(1+\alpha^2)\beta_m{}^2(2\beta_e-\beta_m)D}{\alpha(1+\alpha)\beta_e{}^3H^2} V_4 = 0$$

$$V_1 + V_2 - \frac{(1+\alpha^2)\beta_m{}^3(2\beta_e+\beta_m)}{(1-\alpha)\beta_e{}^3(\beta_e+\beta_m)H^2} V_3$$

$$-\frac{(1+\alpha^2)\beta_m{}^3(2\beta_e-\beta_m)}{(1+\alpha)\beta_e{}^3(\beta_e-\beta_m)H^2} V_4 = 0.$$

Noting that $\alpha > 1$, $\beta_e > \beta_m$, $|\delta| \sim 1$, these can be approximately written

$$V_1 + V_2 + V_3 + V_4 = V \tag{1}$$

$$V_1 - V_2 - jAV_3 - jBV_4 = 0 \tag{2}$$

$$V_1 + V_2 + CV_3 - DV_4 = 0 \tag{3}$$

where A, B, C, D are all large positive numbers.

Now assume that V_3 and V_4 are of the same order of magnitude as V_1, V_2; then from (3)

$$V_3 \cong \frac{D}{C} V_4 \qquad \text{i.e., } V_3 \text{ and } V_4 \text{ have the same sign,}$$

but if this is so, (2) cannot be satisfied, so there is a contradiction unless V_3 and V_4 are small relatively to V_1 and V_2.

Appendix II

The physical significance of the pole exhibited by the effective space charge density at $\gamma = j\beta_e$ in the magnetron amplifier can be demonstrated by setting up the equations of motion:

$$\left(\frac{\partial}{\partial t} + u_0\frac{\partial}{\partial z}\right)\dot{y} = \frac{e}{m}E_y - \omega_c\dot{z}$$

$$\left(\frac{\partial}{\partial t} + u_0\frac{\partial}{\partial z}\right)\dot{z} = \frac{e}{m}E_z + \omega_c\dot{y}.$$

By elimination

$$\left[\left(\frac{\partial}{\partial t} + u_0\frac{\partial}{\partial z}\right)^2 + \omega_c{}^2\right]\dot{z}$$

$$= \frac{e}{m}\left(\frac{\partial}{\partial t} + u_0\frac{\partial}{\partial z}\right)E_z + \omega_c\frac{e}{m}E_y$$

$$\left[\left(\frac{\partial}{\partial t} + u_0\frac{\partial}{\partial z}\right)^2 + \omega_c{}^2\right]\dot{y}$$

$$= \frac{e}{m}\left(\frac{\partial}{\partial t} + u_0\frac{\partial}{\partial z}\right)E_y - \omega_c\frac{e}{m}E_z.$$

Now the equation of continuity is

$$\frac{\partial\rho}{\partial t} + \nabla\cdot\overrightarrow{\rho v} = 0$$

to first order in the ac quantities

$$\frac{\partial\rho}{\partial t} + \nabla\cdot(\overrightarrow{\rho_0 v} + \overrightarrow{u_0\rho}) = 0$$

or

$$\left(\frac{\partial}{\partial t} + u_0\frac{\partial}{\partial z}\right)\rho + \rho_0\left(\frac{\partial\dot{y}}{\partial y} + \frac{\partial\dot{z}}{\partial z}\right) = 0.$$

By differentiating the equations of motion

$$\left[\left(\frac{\partial}{\partial t}+u_0\frac{\partial}{\partial z}\right)^2+\omega_c^2\right]\frac{\partial \dot{z}}{\partial z}$$
$$=\frac{e}{m}\left(\frac{\partial}{\partial t}+u_0\frac{\partial}{\partial z}\right)\frac{\partial E_z}{\partial z}+\omega_c\frac{e}{m}\frac{\partial E_y}{\partial z}$$
$$\left[\left(\frac{\partial}{\partial t}+u_0\frac{\partial}{\partial z}\right)^2+\omega_c^2\right]\frac{\partial \dot{y}}{\partial y}$$
$$=\frac{e}{m}\left(\frac{\partial}{\partial t}+u_0\frac{\partial}{\partial z}\right)\frac{\partial E_y}{\partial y}-\omega_c\frac{e}{m}\frac{\partial E_z}{\partial y}.$$

Adding and substituting from the equation of continuity

$$\left[\left(\frac{\partial}{\partial t}+u_0\frac{\partial}{\partial z}\right)^2+\omega_c^2\right]\left(\frac{\partial}{\partial t}+u_0\frac{\partial}{\partial z}\right)\rho$$
$$=-\frac{e\rho_0}{m}\left(\frac{\partial}{\partial t}+u_0\frac{\partial}{\partial z}\right)\nabla\cdot\vec{E}+\frac{e\rho_0}{m}\vec{\omega}_c\cdot\nabla\times\vec{E}.$$

Now using Poisson's equation, and recognizing $e\rho_0/m\epsilon_0=\omega_p^2$ (where ω_p is the plasma frequency),

$$\left[\left(\frac{\partial}{\partial t}+u_0\frac{\partial}{\partial z}\right)^2+\omega_c^2+\omega_p^2\right]\left(\frac{\partial}{\partial t}+u_0\frac{\partial}{\partial z}\right)\rho$$
$$=\frac{e\rho_0}{m}\vec{\omega}_c\cdot\nabla\times\vec{E}.$$

This equation contains the behavior of the electron gas. Consider the case of a stationary plasma $u_0=0$.

Then

$$\left(\frac{\partial^2}{\partial t^2}+\omega_c^2+\omega_p^2\right)\frac{\partial\rho}{\partial t}=\frac{e\rho_0}{m}\vec{\omega}_c\cdot\nabla\times\vec{E},$$

which means that the plasma "rings" at $\omega=\pm\sqrt{\omega_c^2+\omega_p^2}$ and at $\omega=0$. The zero frequency resonance refers to the drifting motion imparted to the electrons by a dc field. This resonance of course exists also in the nonmagnetic case; it is certainly true that

$$\left(\frac{\partial^2}{\partial t^2}+\omega_p^2\right)\frac{\partial\rho}{\partial t}=0;$$

but in this case the equation is integrable to the second order; that is to say that any forcing term which is not velocity dependent—say, a gravitational potential—would enter as a time derivative.

If $u_0=0$, the implication is that the electron will drift (relative to u_0) in a field that it "sees" as uniform, i.e., a wave traveling with its own speed.

The physical difference between this resonance and the resonance implied by

$$\left[\left(\frac{\partial}{\partial t}+u_0\frac{\partial}{\partial z}\right)^2+\omega_p^2\right]\frac{\partial\rho}{\partial t}=0,$$

which in the idealized case of vanishing ρ_0 coincides with it, is this: in the case of the equation just above, the electrons respond if they "see" the plasma frequency (or, in the magnetic case, $\sqrt{\omega_c^2+\omega_p^2}$); for the third resonance they respond when they "see" a uniform field.

A Wide-Band Voltage-Tunable Oscillator*

J. W. SULLIVAN†

Summary—**A backward-wave oscillator which is electronically tuned over a five to one frequency range is described. Oscillations were observed from 2,600 to 13,300 megacycles with power output up to 50 mw. Equations for starting current and efficiency are derived and compared to experimental results. Other aspects of backward-wave oscillators are also discussed.**

Introduction

THE CHIEF feature of backward-wave oscillator is electronic tuning. It is important to know over how wide a band a backward-wave oscillator (bwo) can be tuned and over how wide a band the circuit can be matched to the output. This depends on the nature of the circuit.

It has been shown that the bifilar helix has an impedance for a backward wave which is many times that of the conventional helix.[1] This allows one to build a voltage-tunable, broadband oscillator which has reasonable dimensions as well as low operating currents. One way to make a bifilar structure is to use two identical wire helixes. The helixes are interwoven as shown in Fig. 1(a). While this structure is easy to make, matching into it is not easy. The high impedance mode for backward oscillations is the plus-minus mode in which the helixes are 180 degrees out of phase. It is difficult to match into this mode since it is necessary to go from waveguide or coax to a balanced line. It was suggested by J. R. Pierce that the above difficulties could be overcome if the structure were made in the following manner: one of the helixes would be formed by grinding a square thread on a rod and the other would be an ordinary wire helix. The wire helix would be screwed into the ground helix and insulated from it. Such a structure is shown in Fig. 1(b). It is evident that this structure is essentially a coaxial line wound into a helix. Part of the outer conductor is cut away to allow space for interaction between the beam and the circuit. Rf power can be removed directly from the circuit through a coaxial line. There

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[1] P. K. Tien, "Bifilar helix for backward wave oscillators," Proc. I.R.E., vol. 42, pp. 1137–1142; July, 1954.

are no frequency sensitive transducers to limit the bandwidth of the match into the structure.

A tube incorporating the circuit of Fig. 1(b) and a hollow electron beam surrounding the structure has been tested. The tube oscillated over a band of 3,500 mc to 13,300 mc when the beam voltage was increased from 73 volts to 3,000 volts. Power outputs of the order

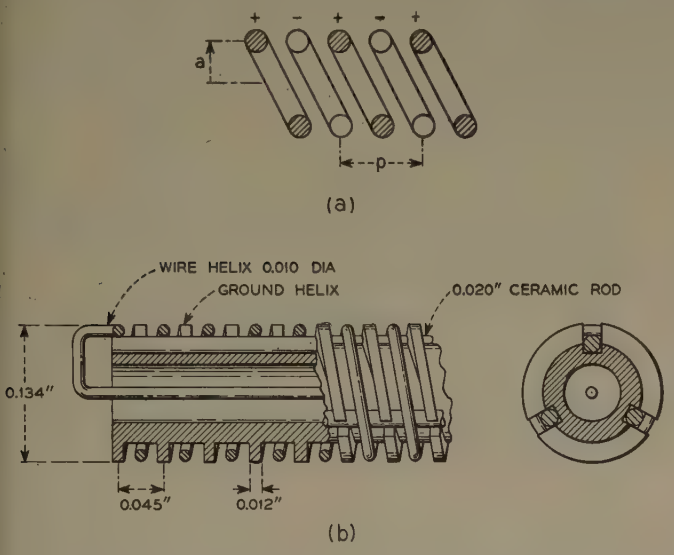


Fig. 1—Two backward-wave structures utilizing bifilar type circuits. A two-wire circuit is shown in (a) and the structure described in the text is depicted in (b).

of 5 milliwatts to 50 milliwatts were obtained with a magnetic focusing field of 470 gauss. When the focusing field was increased to 1,000 gauss, the tube would oscillate down to a beam voltage of 40 volts and a frequency of 2,600 mc. Before describing the experimental results in detail, however, let us write down a few equations which are applicable and will aid in evaluating this tube.

Relation Between Frequency and Voltage

The fundamental equation relating the dc beam velocity to the frequency can be simply stated with the aid of Fig. 2. The beam interacts with the rf wave at

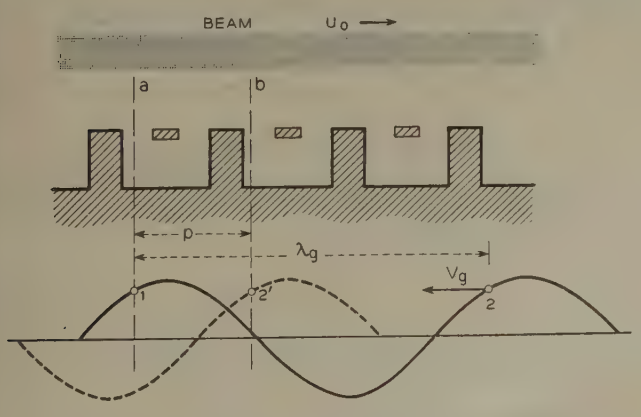


Fig. 2—Circuit used for determining voltage to frequency relationship.

successive gaps, and, if favorable interaction between the beam and the wave is to take place, an electron should see approximately the same phase of the rf wave at each gap. Therefore, we can state that if the phase of the wave is as shown at t_1, when the electron is passing gap a, it must have moved into the phase shown at t_2 when the electron is at the second gap b. In the backward-wave oscillator the wave moves in a direction opposite to the beam, and we then see that the wave must travel a distance $n\lambda_g - p$ in the time $t_2 - t_1$ in order to maintain the correct phase relationship. In this same time the beam will move a distance p.

Thus we have the relationship,

$$\frac{n\lambda_g - p}{v} = \frac{p}{u_0} \qquad n = 1, 2, 3, \cdots, \tag{1}$$

where p is the pitch of the helix, u_0 and v are respectively the electron velocity and the axial velocity of the wave, and $\lambda_g = \lambda_0(v/c)$ is the guide wavelength. Normally, the $n=1$ mode has the highest impedance and consequently oscillations will occur in this mode. If we assume that the wave travels along the helical wire at a constant velocity, (1) can be expressed as

$$V_0(\cot \Psi)^2 = \left[\frac{506ka}{1 - ka}\right]^2, \tag{2}$$

where $ka = 2\pi a/\lambda_0$, a is the helix radius, and Ψ is the pitch angle of the helix ($\cot \Psi = 2\pi a/p$). Eq. (2) is plotted in Fig. 3. From (2) it can be seen that in order

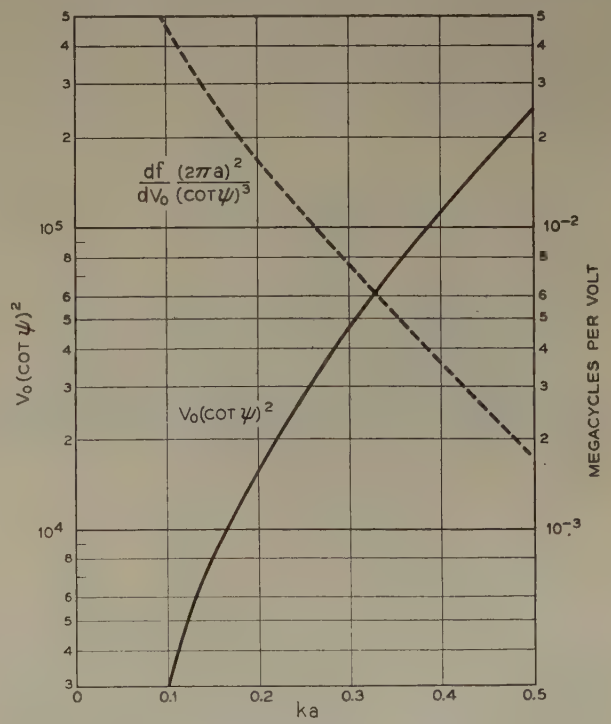


Fig. 3—Normalized tuning curve and derivative for the bwo of Fig. 1(b).

to tune over a n:1 frequency range the voltage range will be given by:

$$\frac{V_{02}}{V_{01}} = n^2\left[\frac{1 - ka}{1 - nka}\right]^2,$$

where V_{02} and V_{01} are respectively the voltages required

for the upper and lower frequencies. Therefore, the minimum voltage range for an n:1 frequency range is n^2, and this is only valid for small values of ka; as one works towards higher values of ka the voltage spread increases very rapidly.

Certain applications for a backward-wave oscillator may require that the frequency, and hence the voltage, be accurately controlled; there may also be occasion to frequency modulate this oscillator. Therefore, it is desirable to find an expression for the incremental change of frequency with voltage. This can be found by differentiating (2). Thus:

$$\frac{df}{dV_0}\frac{(2\pi a)}{(\cot\Psi)^2} = 5.91\times 10^{-4}\frac{(1-ka)^3}{ka}\frac{\text{mcs.-meters}}{\text{volt}}, \quad (4)$$

when a is expressed in meters. Eq. (4) is plotted in Fig. 3. The most significant thing shown by this equation is that the frequency change per volt becomes much higher as ka decreases. For instance, at $ka=0.1$ the frequency changes with voltage about thirty times as fast as it does at $ka=0.5$. The design of a tube for wide-band operation will then be a compromise between large voltage changes and small dimensions.

The foregoing equations assume the wave travels along the helical wire at the velocity of light. This characteristic may be altered somewhat by dielectric loading. The amount of dielectric loading on any helix structure can either be calculated or measured by cold tests. Using this loading factor and the physical dimensions of the structure, one has the information necessary to plot frequency as a function of beam voltage.

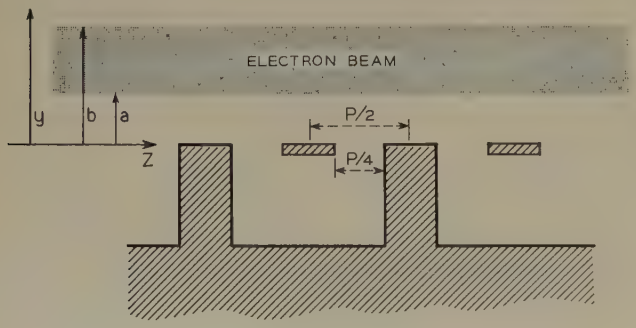


Fig. 4—Configuration used in calculating the circuit impedance of the bwo.

Circuit Impedance and Starting Current

In evaluating a given rf structure for use as a bwo, one must consider at least two importance factors. First, the voltage tuning curve mentioned in the preceding section, and second, the impedance of the structure as seen by the electron beam. We shall discuss this impedance in terms of the familiar parameter $E^2/_2\beta^2P$ where E is the peak longitudinal electric field at the beam position, β is given by ω/u_0, and P is the rf power carried by the circuit.

If the wire helix of the circuit in Fig. 1(b) is replaced by a tape helix, the resulting circuit will look like that shown in Fig. 4. The impedance of this structure can be approximated as follows. At a given value of y, the magnitude of the traveling wave field produced by a voltage, V, acting across the gap can be taken as

$$E = 2M\frac{V}{p}e^{-\gamma y}, \quad (5)$$

where $M=(\sin\theta_g/2)/\theta_g/2$ is the gap modulation coefficient, γ is the transverse propagation constant ($\gamma\approx\beta$), and $\theta_g=\beta p/4$ is the gap transit angle.[2]

The power carried by a circuit with a characteristic impedance Z_0 is

$$P = \frac{V^2}{2Z_0}. \quad (6)$$

By using (5) and (6), a simple expression for impedance can be obtained.

$$K(y) = \frac{M^2Z_0}{4\theta_g{}^2}e^{-2\gamma y}. \quad (7)$$

The gap angle, θ_g, can be expressed as

$$\theta_g = \frac{\pi}{2}(1-ka). \quad (8)$$

This becomes apparent when two cases are considered. At low frequencies where $ka\sim 0$, the voltage phase change per section is 2π and consequently $\theta_g=\pi/2$. When $ka=0.5$, the phase change per section is π and $\theta_g=\pi/4$.

Eq. (7) gives impedance as a function of frequency at a given value of y. Suppose a beam extending from $a\leqq y\leqq a+t$ is sent along the structure. If the current density in the beam is uniform, the impedance seen by the beam is

$$K = \frac{M^2Z_0}{4\theta_g{}^2}\int_a^{a+t}\frac{e^{-2\gamma y}}{t}dy = \frac{M^2Z_0}{32\theta_g{}^3}\frac{e^{-8\theta_g a/p}\left[1-e^{-8\theta_g t/p}\right]}{(t/p)}, \quad (9)$$

since $\theta_g=\gamma p/4$. If $t\gg p$, (9) reduces to

$$K = \frac{M^2Z_0e^{-8\theta_g a/p}}{32\theta_g{}^3(t/p)}. \quad (10)$$

Once the circuit impedance has been determined it is a simple matter to calculate the starting current. In a lossless circuit which is short in terms of plasma wavelengths, oscillations occur when

$$CN = 0.314, \quad (11)$$

where C is the gain parameter and N is the length of the active circuit measured in wavelengths.[3,4] When

[2] J. R. Pierce, "Traveling-Wave Tubes," D. Van Nostrand Company, New York, N. Y., chap. 4, pp. 73–74; 1950.
[3] R. Kompfner and N. T. Williams, "Backward wave tubes," Proc. I.R.E., vol. 41, pp. 1602–1611; November, 1953.
[4] H. Heffner, "Analysis of the backward wave traveling wave tube," Proc. I.R.E., vol. 42, pp. 930–937; June, 1954.

pace charge can not be ignored, the value of CN for tarting can be found from Fig. 5. The factor Q/N is efined

$$Q/N = \frac{1.87 \times 10^5 \lambda_0^3 V_0}{KLA}\left(\frac{\omega_q}{\omega_p}\right)^2, \tag{12}$$

where Q is the space charge parameter, L is the active ength of the circuit, A is the beam area, and ω_q/ω_p is

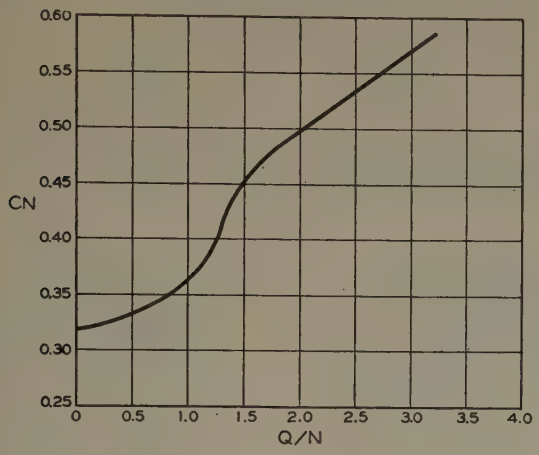


Fig. 5—Zero loss starting conditions for a backward-wave oscillator.

he space charge reduction factor which is calculated in he Appendix.[5] When (11) is solved for starting current, I_{0s}, there result

$$I_{0s} = \frac{.124V_0}{KN^3}. \tag{13}$$

Where ω_p is the plasma frequency,

$$\omega_p = 1.82 \times 10^8 \left(\frac{J_0}{\sqrt{V_0}}\right)^{1/2}. \tag{15}$$

The factor (ω_q/ω_p) represents a reduction in plasma frequency due to the proximity of conducting media and finite beam dimensions. The term $1/2\gamma t$ is a reduction due to the fact that fields are not constant across the beam. This correction holds only when $\gamma t \gg 1$. If this is not the case, $1/2\gamma t$ should be replaced by $(1-e^{-2\gamma t})/2\gamma t$. This factor as well as ω_q/ω_p is discussed further in the appendix.

The output power with (14) and (15) becomes

$$P_{\text{out}} = \frac{1.82 \times 10^8}{A^{1/2}\omega}\left(\frac{\omega_q}{\omega_p}\right)\left(\frac{1}{2\gamma t}\right) I_0^{3/2} V_0^{3/4}, \tag{16}$$

where A is the cross-sectional area of the beam.

Experimental Results

The results described below were observed on a tube similar to that shown in Fig. 6. The rf circuit is supported at the collector end; the wire helix is connected to the center conductor of a coaxial line which leaves the tube through a kovar to glass coaxial seal. The characteristic impedance of the rf circuit was determined by electrolytic tank measurements and was found to be 80 ohms. Since most coaxial measuring equipment has an impedance of 50 ohms, it is necessary to transfer from 80 ohms at the input to the wire helix to 50 ohms at the output of the tube. This was accomplished by tapering the diameter of the coaxial center conductor from .014 inch at the gun end of the line to 0.023

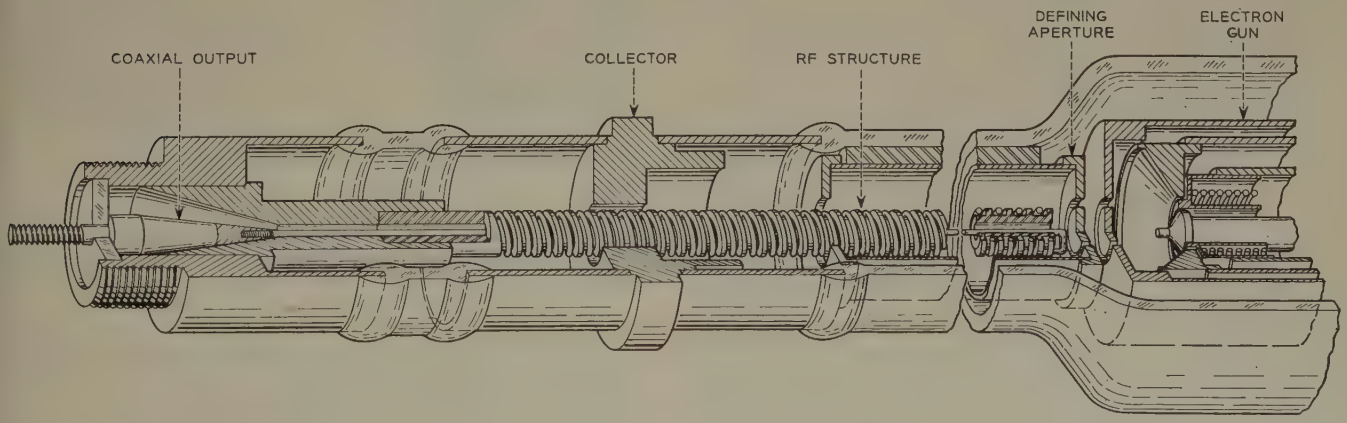


Fig. 6—Schematic drawing of the oscillator described in the text.

Efficiency and Output Power

The limiting efficiency of a bwo has been discussed in he following form:[6]

$$\text{Efficiency} = \frac{\omega_p}{\omega}\left(\frac{\omega_q}{\omega_p}\right)\frac{1}{2\gamma t}. \tag{14}$$

inch at the collector end. Before the coaxial line leaves the tube, both the inner and the outer conductors are tapered up to standard dimensions for 50-ohm coaxial fittings. At the collector end the circuit is terminated by spraying aquadag on the ceramic rods which separate the helixes. An annular beam surrounds the circuit, and the beam is collected before the loss section. Beam current is controlled changing the voltage on the first anode and the voltage on the circuit and second anode is swept through the required range.

[5] Since Q/N is independent of current, it is more convenient to lot CN as a function of this parameter instead of $\omega_p L/u_0$ as was done n reference 4.

[6] This was presented by Grow and Watkins at the Electron Tube Conference at Stanford Univ., Calif.: June, 1953.

From cold tests it was found that the three ceramic support rods reduced the wave velocity to 0.91 *c*. Furthermore, it appeared that this loading factor was independent of frequency. This, together with the tube dimensions, enables one to calculate the tuning curve shown in Fig. 7. Experimental points fell on this curve indicating a very good agreement with theory. Oscillations were observed continuously from 4,000 to 13,300 mcs as the voltage was swept from 100 to 3,000v. This was with a field of 470 gauss. By increasing the field to 1,000 gauss, the tube would oscillate down to 2,600 mcs (40v).

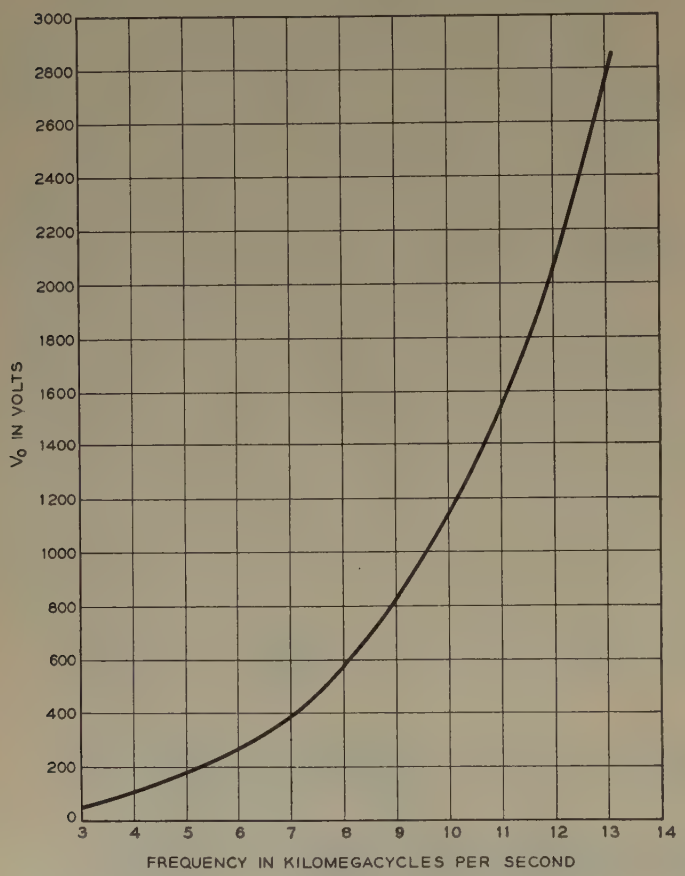


Fig. 7—Tuning curve for the circuit shown in Fig. 1(b).

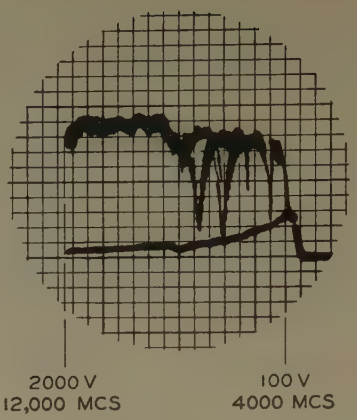


Fig. 8—Oscilloscope picture of output power as the beam voltage i swept from 100 to 2,000 volts. Height above the curved base lin represents output from the detector crystal.

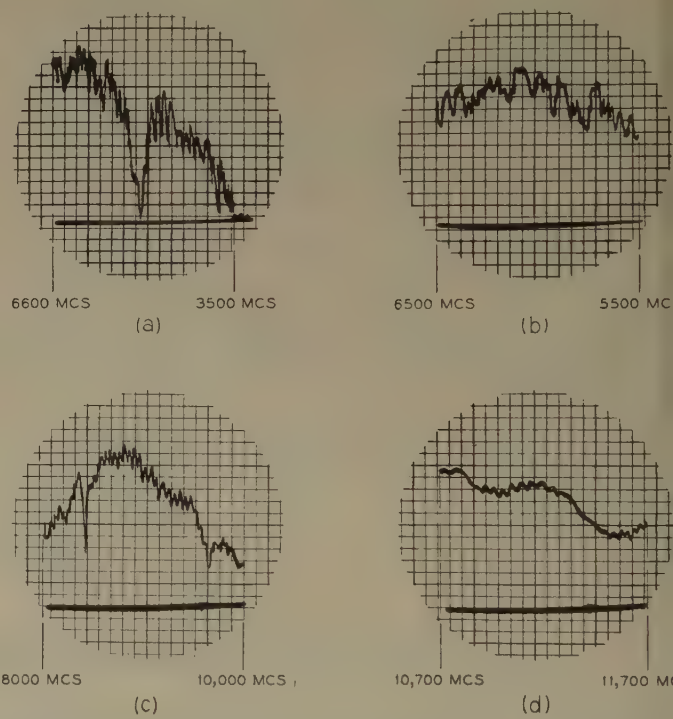


Fig. 9—Oscilloscope pictures of output power over smaller frequenc bands as observed from crystal detectors in appropriate size wav guide.

Fig. 8 shows a photograph of the output of the oscillator vs frequency. The voltage is swept from 100 to 2,100v and the frequency goes from 4,000 to 12,100 mcs. The output is detected by a coaxial crystal and the height above the curved base line is a measure of output power. The curved base line is because the helixes are not connected together for dc. There was, consequently, some dc current flowing through the crystal. It should be borne in mind that the electrical properties of the crystals and their associated equipment is tested as well as the tube characteristics. One of the holes in the output curve shown in Fig. 8 could be moved by twisting a coaxial fitting. Figs. 9(a), (b), (c), and (d) show output vs frequency for four smaller frequency bands. The hole occurring at about 5,000 mc (in Fig. 9(a)) is believed to be associated with the crystal mount

Since the beam is very close to the rf circuit, it i almost impossible to avoid intercepting current on thi structure. Consequently, the current reaching the co lector may not represent all the active current. Fig. 1 shows (at start of oscillation) both collector current an collector plus intercepted current as functions of fre quency the true starting current should lie somewher between these two values. Also shown is starting cu rent as calculated from (13). The agreement is fairl good (at least qualitatively).

Fig. 11 shows output power vs frequency for a con stant beam current of 10 milliamperes. The theoretica curve calculated from (16) is also shown. It is surprisin that there is such a close agreement between the ob

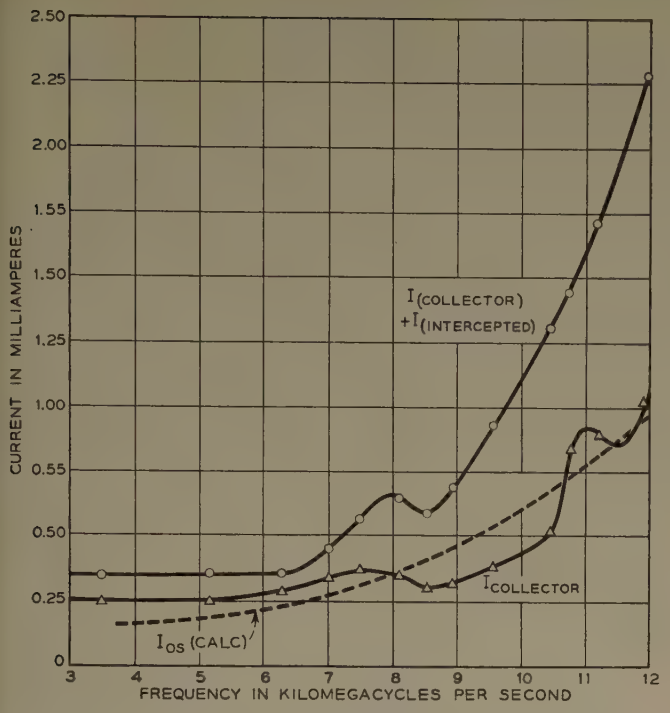


Fig. 10—Computed and observed starting current as a function of frequency.

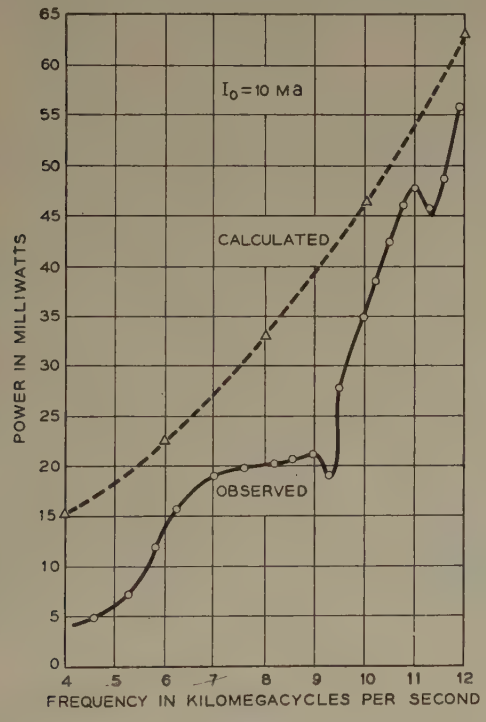


Fig. 11—Calculated and observed output power as a function of frequency.

served and calculated output power. It is also seen that the output power varies only 10:1 in a 3:1 frequency band for constant beam current.

Frequency Pulling

The effect of an external load upon the frequency of an oscillator is commonly called "frequency pulling." The standing wave into this tube was varied from unity to infinity and no change in frequency was observed. However, as beam current is increased (with beam voltage held constant), frequency decreases. Fig. 12 shows that the frequency decreased 45 mcs (out of 10,000 mcs) as the beam current is increased from starting current to three times starting current.

Conclusion

The results given above show a fair degree of agreement with those predicted by theory. However, the gratifying thing about this tube is that it shows one tube is capable of oscillating continuously over a five-to-one frequency band. Another thing which becomes evident is the need for broadband components to be used with such a tube. A good crystal detector would be especially useful.

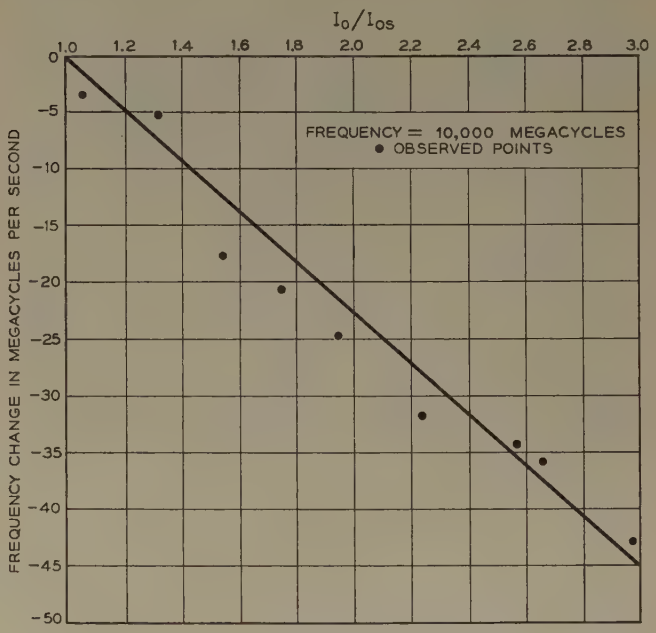


Fig. 12—Frequency change as a function of beam current with beam voltage held constant.

Acknowledgment

The author wishes to express his appreciation to the members of the Electronics Research Department of Bell Telephone Laboratories for their many helpful suggestions. Thanks are especially due to J. R. Pierce and C. F. Quate for their encouragement throughout this investigation.

The parts for the experimental tubes were prepared under the direction of P. Hannes, and the tubes were assembled by E. J. Flannery.

Appendix

When the beam is not infinite, a space charge reduction factor ω_q/ω_p must be found. This reduction factor will depend upon the constants of the beam and its surroundings.

Consider the circuit shown in Fig. 13. A conducting plane at $y=0$ represents the circuit. The beam is infinite in the x-direction and exists for $a \leqq y \leqq b$. Propagation

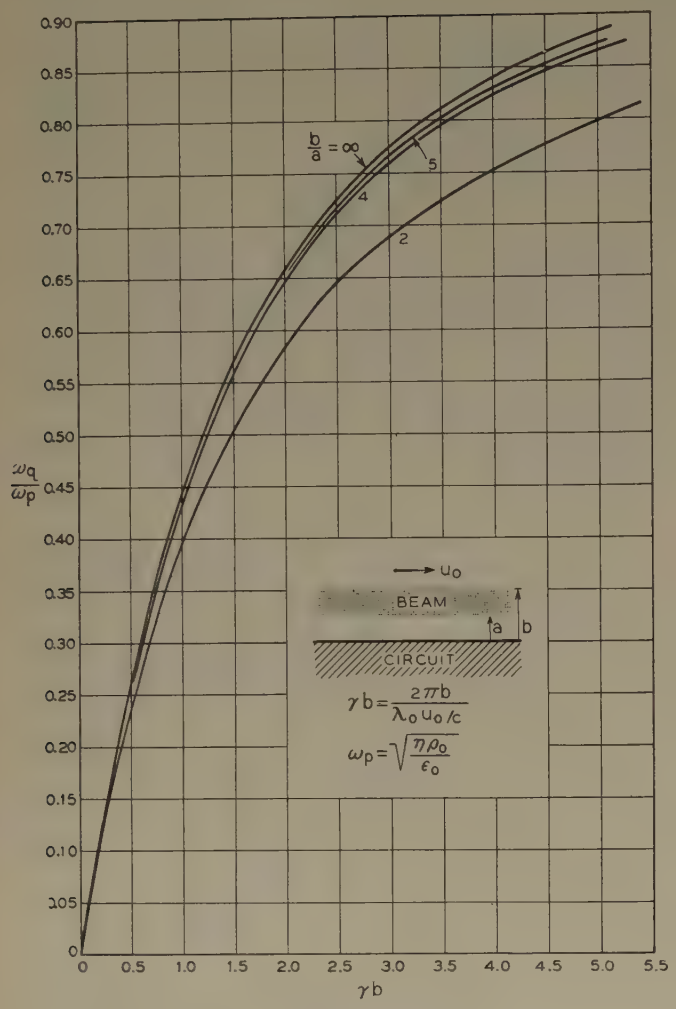


Fig. 13—Plasma frequency reduction vs γb with b/z as a parameter.

is along the z-axis as $e^{j(\omega t-\gamma z)}$. Furthermore only z-directed forces will be considered. ω_q is defined such that

$$\gamma \pm \frac{\omega \pm \omega_q}{u_0}. \tag{17}$$

Starting with Maxwell's equations and neglecting second order terms, one can derive the following relationship for E_z:

$$\nabla_t^2 E_z = -\gamma^2\left(\frac{\omega_p^2}{\omega_q^2} - 1\right)E_z. \tag{18}$$

For the geometry being considered,

$$\nabla_T^2 = \frac{\partial^2}{\partial y^2}.$$

Appropriate solutions for (18) are for $a \geqq y \geqq 0$.

$$E_{zI} = A \sinh \gamma y + B \cosh \gamma y; \tag{19}$$

for $b \geqq y \geqq a$

$$E_{zII} = C \sin Ty + D \cos Ty; \tag{20}$$

and for $b \leqq y$

$$E_{zIII} = Fe^{-\gamma y}. \tag{21}$$

There are five boundary conditions which can be imposed; they are

at

$$y = 0, \qquad E_{zI} = 0$$

at

$$y = a, \qquad E_{zI} = E_{zII}$$

and

$$\frac{dE_{zI}}{dy} = \frac{dE_{zII}}{dy}$$

and at

$$y = b$$
$$E_{zII} = E_{zIII}$$

and

$$\frac{dE_{zII}}{dy} = \frac{dE_{zIII}}{dy}.$$

The above field equations and boundary conditions give an expression for the propagation constant. This is

$$\frac{\gamma^2}{T_2} + \frac{\gamma}{T}\left[1 + \tanh \gamma a\right]\cot T(b-a) - \tanh(\gamma a) = 0. \tag{22}$$

Eq. (22) can be solved for γ/T as a function of γb for various ratios of b/a. Then by the use of (18) the relationship

$$\frac{\omega_q}{\omega_p} = \left[\frac{1}{1 + \dfrac{T^2}{\gamma^2}}\right]^{1/2} \tag{23}$$

can be derived. Thus ω_q/ω_p can be plotted as a function of γb. This is done in Fig. 13.

There is another factor which reduces the efficiency below that given by ω_q/ω. The field is not uniform across the beam. That is, if the current branches to I_0 on the inside of the beam, the current across the beam will be given by

$$i = I_0 e^{-\gamma y}. \tag{24}$$

Also the field across the beam is

$$E = E_0 e^{-\gamma y}. \tag{25}$$

The efficiency as given by ω_q/ω is based on an output power which is proportional to (tI_0E_0) where $t = b - a$ is the beam thickness.

$$\int_0^t \frac{e^{-2\gamma y}}{t}\,dy = \frac{1 - e^{-2\gamma t}}{2\gamma t}. \tag{26}$$

If $\gamma t > 1$, $e^{-2\gamma t} < .135$, and consequently the output power will be reduced by approximately $1/2\gamma t$.

The complete expression for output power is:

$$P_{\text{out}} = \left(\frac{\omega_q}{\omega_p}\right)\left(\frac{\omega_p}{\omega}\right)\frac{I_0 V_0}{2\gamma t}. \tag{27}$$

List of Symbols

p = pitch of the helix
a = mean radius of the helix
ψ = helix pitch angle;

$$\cot \psi = \frac{2\pi a}{p}$$

λ_0 = free space wavelength
c = velocity of light
V_0 = dc beam voltage
ka = circumference of helix in wavelengths;

$$ka = \frac{2\pi a}{\lambda_0}$$

f = frequency
γ = transverse propagation constant; field decay as $e^{-\gamma y}$
M = gap modulation coefficient;

$$M = \frac{\sin \frac{\theta_g}{2}}{\frac{\theta_g}{2}}$$

θ_g = gap transit angle;

$$\theta_g = \frac{\gamma p}{4} = \frac{\pi}{2}(1 - ka)$$

P = power
z_0 = characteristic impedance of a circuit
E = electric field
K = impedance parameter;

$$K = \frac{E^2}{\beta^2 P}$$

C = gain parameter;

$$C^3 = \frac{I_0 K}{4V_0}$$

N = length of circuit in wavelengths;

$$N = \frac{L}{\lambda_0 \frac{u_0}{c}}$$

L = length of circuit
u_0 = beam velocity
ω_p = plasma frequency;

$$\omega_p = \left(\frac{\eta J_0}{\epsilon u_0}\right)^{1/2}$$

ω_q = reduced plasma frequency
ω = angular frequency
R = efficiency parameter
Q = space charge parameter
δ = incremental propagation constant; $\delta = x + jy$.

Synthesis of Constant-Time-Delay Ladder Networks Using Bessel Polynomials*

LEO STORCH†

Summary—A realizable, low-pass, minimum-phase transfer function is constructed which produces "maximally flat" time delay as a function of frequency, and which is associated with an impulse response that resembles closely a Gaussian curve properly delayed with respect to $t=0$. The delay and loss characteristics are particularly easy to evaluate, since this transfer function can be expressed in terms of conveniently tabulated functions.

It is shown how to realize this transfer function as the transfer impedance of a ladder network containing only low-Q elements, and the response of this network is put into perspective to those of conventionally designed delay networks.

Also, a class of nonminimum-phase transfer functions is generated from the above transfer functions in order to offer added flexibility. The delay and loss characteristics of the nonminimum-phase transfer functions are described, but their network realization is not considered in this paper.

Introduction

THE DESIGN of lumped-constant delay networks has been dominated up to now by image-parameter filters, with particular attention devoted to linearizing their phase-angle versus frequency curve in the case of the m-derived or more complex sections. This type of design is characterized by the assumption that all reactors are lossless and by the cascade-connection of several alike sections in order to obtain a desired over-all bandwidth-delay product. The low-pass filters usually turn out to be of the bridged-T type with negative mutual inductance, a capacitor in the bridge arm, and maybe added shunt capacitors across input and output terminals. They appear to have poor pulse and square-wave response on account of an additional high-pass band with an uncontrolled phase characteristic. A few references[1–4] will orient the reader in this field, with no slight intended to the numerous papers not mentioned here.

With the advent of modern network synthesis, a more flexible approach has become feasible. One can afford to concentrate on the over-all transfer function and to synthesize subsequently a suitable network realization, rather than being limited to adapting one of the relatively small number of tractable network sections to a specific application and achieving the desired over-all effect by cascading several sections on an image-impedance basis.

In Part I of this paper, a realizable, low-pass, minimum-phase approximation of the ideal distortionless delay operator e^{-pt_0} will be discussed ("p" is the complex-frequency variable $\sigma+j\omega$; "t_0" is the fixed delay time). It is a "best" approximation in the sense that it produces "maximally flat" delay in the frequency domain and, accordingly, a rather symmetrical impulse response in the time domain, which is centered about $t=t_0$ and approaches the Gaussian curve as the degree of the approximation is increased. This transfer function will be synthesized in Part II of the paper in the form of a passive low-pass ladder network. A resistor will be associated with each reactor, which can be chosen to correspond to a rather low Q at the reference frequency $\omega_0=1/t_0$. These resistive elements are taken into account as part of the synthesis procedure, so that their presence causes merely a fixed loss but no distortion of the desired transfer function. Subsequently, the circuit for the ninth-degree approximation (Fig. 6) is used as an illustration of the method and oscillograms of its square-wave response are compared in Fig. 7 with those of a conventional bridged-T network of comparable delay-bandwidth product, but approximately twelve times as much coil weight and volume (Fig. 8).

Also in Part I, the primary minimum-phase transfer functions are combined in order to generate groups of nonminimum-phase transfer functions, which possess maximally flat delay but modified loss characteristics and impulse responses. Their loss and delay characteristics are specified, but the synthesis of the corresponding networks falls outside the scope of this paper. Both types of transfer functions are closely related to Bessel polynomials, which have been studied in the recent mathematical literature. In addition to their desirable properties as approximants of e^{pt_0}, they offer the convenience of effortless evaluation in most cases, since they are expressible in terms of the comprehensively tabulated spherical Bessel functions.

Part I

A Realizable Low-Pass Approximation of the Ideal Delay Transfer Function e^{-pt_0}

An ideal delay network would produce an output which is exactly like the input but shifted along the time axis in the positive direction by t_0, the desired delay time. A network which had the property that $f_{out}(t)=f_{in}(t-t_0)$ would have to possess e^{-pt_0} as a transfer function[5] and a δ-function at $t=t_0$ as the response to a

* Decimal classification: R143.4. Published also in the 1954 IRE Convention Record, Part 2—Circuit Theory.

† Hughes Aircraft Co., Culver City, Calif.

[1] L. Batchelder, U. S. Patent 2,250,461; 1941.

[2] M. H. Herb, C. W. Horton, and F. B. Jones, "On the design of networks for constant time delay," *Jour. Appl. Phys.*, vol. 20, pp. 616–620; June, 1949.

[3] C. F. Floyd, R. L. Corke, and H. Lewis, "The design of linear phase low-pass filters," *Proc. IEE*, part IIIA, pp. 777–787; 1952.

[4] C. M. Wallis, "Design of low-frequency constant time delay lines," *Trans. AIEE*, part I, pp. 135–140; 1952.

[5] M. F. Gardner and J. L. Barnes, "Transients in Linear Systems," John Wiley and Sons, Inc., New York, N. Y., vol. I, p. 236; 1942.

δ-function input at $t=0$. Relaxing this requirement somewhat, an ideal distortionless delay network is characterized by the transcendental transfer function ke^{-pt_0}, since the positive real constant "k" is responsible merely for a fixed gain or loss without introducing distortion in its usual meaning.[6] The prominent property of a distortionless delay network is the linear relationship of phase-angle versus real frequency ($p=j\omega$) for the infinite range $0<\omega<\infty$.

But a finite, linear, passive, lumped-constant network is never associated with a transfer function outside the class of rational functions of $p=\sigma+j\omega$ and expressible as $T(p)=P(p)/Q(p)$, such that $P(p)$ and $Q(p)$ are two polynomials in "p" with real coefficients, the roots of $Q(p)$ lie in the left half of the p-plane, and $P(p)$ does not exceed $Q(p)$ in degree.[7] Further restrictions on $P(p)$ and $Q(p)$ may be imposed if a particular type of network is desired, e.g., unbalanced, lc, rc. The transfer function may be represented physically by a dimensionless voltage or current ratio, a transfer impedance, or a transfer admittance. Although roots of $Q(p)$ on the real-frequency axis, $p=j\omega$, are in general permissible, all roots will be required to lie *inside* the left half of the p-plane in view of the presence of resistive elements in the network.[8] In other words, $Q(p)$ is required to be a Hurwitz polynomial.

4. *Test for Stability (Hurwitz Polynomials)*

In the establishment of a realizable transfer function suitable as an approximation of the ideal delay operator, the most stringent restraint that must be imposed is the Hurwitz property of the denominator $Q(p)$, which determines the natural modes of the network.

Let $Q_N(p)$ be an N'th degree polynomial in "p" with real coefficients, and let it be written as

$$Q_N(p) = G_N(p) + H_N(p) = \sum_{r=0}^{N} a_r p^r, \tag{1}$$

such that $G_N(p)$ contains only the terms with even powers in "p" and $H_N(p)$ only the odd ones. The necessary and sufficient condition that $Q_N(p)$ be a Hurwitz polynomial is simply that the coefficients of all N terms must be positive when $G_N(p)/H_N(p)$ is expanded as a continued fraction in "$1/p$" or "p."[9]

An analogous statement can be phrased in terms of network theory:[10] if $Q_N(p)$ is a Hurwitz polynomial with real coefficients, then $G_N(p)/H_N(p)$ must have all the properties of a purely reactive driving-point immittance (impedance or admittance) function, and the converse is also true. The theorem can be linked directly to the preceding statement, since a continued fraction in "p" or "$1/p$" with positive real coefficients can be diagrammed readily as the driving-point immittance of a purely reactive ladder network with alternating capacitive and inductive arms (Cauer's canonical representations[11]).

This attribute of Hurwitz polynomials offers a convenient but cogent method for enforcing the Hurwitz property of the denominator of the transfer function, and forms the principal ingredient, moreover, of the subsequent synthesis procedure.

B. A Transfer Function for "Maximally Flat" Low-Pass Delay Networks.

For simplicity's sake, let t_0 be absorbed in "p" as a scale-factor, so that the ideal delay operator may be written as

$$e^{-p} = \frac{1}{\cosh p + \sinh p}. \tag{2}$$

Let the denominator of (2) be approximated by an N'th degree polynomial, establishing the form of the transfer function as

$$T_N(p) = \frac{a_0}{Q_N(p)} = \frac{a_0}{\sum_{r=0}^{N} a_r p^r} = \frac{a_0}{G_N(p) + H_N(p)}. \tag{3}$$

In order to maintain control over the Hurwitz property of $Q_N(p)$ from the beginning, let the approximation be initiated by expanding $\cosh p/\sinh p$ of (2) in an infinite continued fraction. Making use of the continued-fraction expansion of $J_{\gamma-1}(z)/J_\gamma(z)$ as given by Watson:[12]

$$\frac{\cosh p}{\sinh p} = \frac{J_{-1/2}(-jp)}{j\cdot J_{1/2}(-jp)} = \frac{1}{p} + \cfrac{1}{\cfrac{3}{p} + \cfrac{1}{\cfrac{5}{p} + \cfrac{1}{\cfrac{7}{p} + \cdots}}} \tag{4}$$

If the expansion (4) is truncated at the N'th term, a finite continued fraction

$$\frac{G_N(p)}{H_N(p)} = \frac{1}{p} + \cfrac{1}{\cfrac{3}{p} + \cfrac{1}{\cfrac{5}{p} + \cdots + \cfrac{1}{\cfrac{2N-1}{p}}}} \tag{5}$$

[6] M. J. Di Toro, "Phase and amplitude distortion in linear networks," Proc. I.R.E., vol. 36, pp. 24–36; January, 1948.

[7] H. W. Bode, "Network Analysis and Feedback Amplifier Design," D. Van Nostrand, Inc., New York, N. Y., chaps. I and XI; 1945.

[8] A. E. Guillemin, "Modern Methods of Network Synthesis," *Advances in Electronics*, Academic Press, New York, N. Y., vol. III, p. 275; 1951.

[9] H. S. Wall, "Polynomials whose zeros have negative real parts," *The Amer. Math. Month.*, vol. 52, pp. 308–322; June–July, 1945.

[10] A. E. Guillemin, "The Mathematics of Circuit Analysis," John Wiley and Sons, Inc., New York, N. Y., pp. 395–409; 1949.

[10] W. Cauer, "Frequenzweichen konstanten Betriebswiderstandes," *Elek. Nach. Tech.*, vol. 16, p. 116; 1939.

[11] A. E. Guillemin, "Communication Networks," John Wiley and Sons, Inc., New York, N. Y., vol. II, pp. 198–202; 1935.

[12] G. N. Watson, "A Treatise on the Theory of Bessel Functions," Cambridge University Press, p. 303; 1952.

is obtained, which approximates $\coth p$ in the vicinity of $p=0$. In view of the fact that $\beta_k=1+2k$, for $k=0, 1, 2, \cdots, (N-1)$, all the β_k's are nonzero and positive. Therefore, when the process is reversed and the terms of (5) are combined to form a $Q_N(p)$ with specific coefficients a_r, which are functions of N, this $Q_N(p)$ must be a Hurwitz polynomial for any finite N. Besides assuring the Hurwitz property of the denominator of the transfer function (3), focusing the approximation on the ratio of the even to the odd part, $G_N(p)/H_N(p)$, is particularly appropriate for another important reason. This expression exclusively determines the phase angle and, implicitly, the delay characteristic of (3) for real frequencies $p=j\omega t_0$.

The general form of $Q_N(p)$ is obtained most expeditiously by expressing (5) first as the ratio of two polynomials in $1/p$, i.e.,

$$G_N(1/p)=\frac{G_N(p)}{p^N} \quad \text{and} \quad H_N(1/p)=\frac{H_N(p)}{p^N}.$$

These can be obtained by means of standard recursion formulas for continued fractions,[13] which are of the same form for $G_k(1/p)$ and $H_k(1/p)$, $k=1, 2, \cdots, N$. As applied to (5), the recursion formula for $Q_k(1/p)=G_k(1/p)+H_k(1/p)$ is:

$$Q_k(1/p)=\frac{2k-1}{p}Q_{k-1}(1/p)+Q_{k-2}(1/p), \qquad (6)$$

with initial values $Q_{-1}=1$, $Q_0=1$. The recursion formula (6), with its associated initial values, is sufficient to identify $Q_N(1/p)$ as the Bessel polynomial $y_N(1/p)$:[14]

$$Q_N(1/p)=y_N(1/p)=\sum_{r=0}^{N}\frac{(N+r)!}{(N-r)!r!(2p)^r}. \qquad (7)$$

Consequently:

$$T_N(p)=\frac{a_0}{Q_N(p)}=\frac{a_0}{p^NQ_N(1/p)}=\frac{a_0}{\sum_{r=0}^{N}a_rp^r}, \qquad (8)$$

where the coefficients are the positive integers

$$a_r=\frac{(2N-r)!}{2^{N-r}r!(N-r)!}=\binom{2N-r}{r}\cdot[1\cdot3\cdot5\cdot\cdots\cdot(2N-1-2r)], \qquad (9)$$

and in particular

$$a_0=\frac{(2N)!}{2^N(N)!}=1\cdot3\cdot5\cdot7\cdot\cdots\cdot(2N-1).$$

It may be reassuring to compare $Q_N(p)/a_0$ with the McLaurin series of e^p, i.e., $e^p=\sum_{r=0}^{\infty}1/r!\,p^r$. If a_r in (9) is expanded, then it is seen that $(r>1)$

$$\frac{a_r}{a_0}=\frac{(N-1)(N-2)(N-3)\cdots(N-r+1)}{r!\left(N-\frac{1}{2}\right)(N-1)\left(N-\frac{3}{2}\right)\cdots\left(N-\frac{r-1}{2}\right)}.$$

Since both the numerator and the denominator contain "r" terms, $(a_r/a_0)\to(1/r!)$ as $N\to\infty$, so that $Q_N(p)/a_0$ equals the McLaurin series of e^p in the limit. Two questions are likely to arise in the reader's mind: (a) why not use the straightforward McLaurin series in the first place; (b) how would the properties of a network synthesized on the basis of the McLaurin series differ from those of a network based on (8) and (9). These questions can be forestalled by the simple statement that a polynomial formed from the first N terms of the McLaurin series of e^p is not a Hurwitz polynomial when N is larger than 4.[15] Therefore, it is not eligible as the denominator of a realizable transfer function for N larger than 4, which includes the majority of applications.

C. Delay and Loss Properties of $T_N(p)$

The study of the amplitude and phase characteristics of the transfer function (8) and the choice of a suitable N are greatly simplified by taking advantage of the relationship between Bessel polynomials and the spherical Bessel functions

$$\sqrt{\frac{\pi}{2x}}J_{\pm(\gamma+1/2)}(x),$$

for which reasonably comprehensive tables have been published.[16] In this context, it is necessary to deal with real frequencies, i.e., $p=j\omega t_0=ju$. Inasmuch as it can be shown[17] that

$$y_N(1/ju)=j^{-N}\sqrt{\frac{\pi u}{2}}e^{iu}[(-1)^NJ_{-N-1/2}(u)-j\cdot J_{N+1/2}(u)] \qquad (10)$$

and since $Q_N(ju)=(ju)^Ny_N(1/ju)$ by (7) and (8), therefore,

$$T_N(ju)=\frac{a_0\cdot e^{-iu}}{u^{N+1}\sqrt{\frac{\pi}{2u}}[(-1)^NJ_{-N-1/2}(u)-j\cdot J_{N+1/2}(u)]} \qquad (11)$$

where $a_0=(2N)!/2^NN!$, as in (9).

One is now in the pleasant position of being able to study and explore the properties of the transfer function merely by looking up values in published tables and multiplying them by u^{N+1}, rather than having to devote time and energy to the numerical evaluation of N'th degree polynomials. Expression (11) has the additional merit of displaying explicitly the distortion with respect to the ideal delay operator e^{-iu}. From

[13] O. Perron, "Die Lehre von den Kettenbruechen," Teubner, Berlin, Ger., p. 5; 1929.

[14] H. L. Krall and O. Fink, "A new class of orthogonal polynomials: the Bessel polynomials," *Trans. Amer. Math. Soc.*, vol. 65, pp. 100–115 (see p. 101); January, 1949.

[15] K. E. Iverson, "The Zeros of the Partial Sums of e^z," *MTAC*, vol. VII, pp. 165–168; July, 1953.

[16] NBS, Mathematical Tables Project, "Tables of Spherical Bessel Functions," 2 vols., Columbia University Press, New York, N. Y.; 1947.

[17] H. L. Krall and O. Fink, *op. cit.*, p. 103.

$$T_N(ju) = Ae^{-i\theta} = (\text{Distortion factor}) \cdot e^{-iu} = (Ae^{i\epsilon}) \cdot e^{-iu}, \quad (12)$$

it follows that phase distortion as measured by the angle of lag "ϵ" with respect to the ideal value of "u" radians is given by

$$\epsilon = \tan^{-1} \frac{J_{N+1/2}(u)}{(-1)^N J_{-N-1/2}(u)} \quad (13)$$

and amplitude distortion is given by

$$A = \frac{a_0}{u^{N+1}\left\{\frac{\pi}{2u}\left[J^2_{-N-1/2}(u) + J^2_{N+1/2}(u)\right]\right\}^{1/2}}. \quad (14)$$

The loss $L = -20 \log A$ db tends to the Gaussian form

$$L \to \frac{10u^2}{(2N-1) \ln 10} \, db$$

with increasing N, which corresponds to an approximate 3 db bandwidth of $u_{3db} \cong \sqrt{(2N-1) \ln 2}$. This approximation is serviceable for values of N larger than 3.[18]

There are two methods of defining delay that are in common use:[2] phase delay $t_{ph} = \theta/\omega$, and group delay $t_g = d\theta/d\omega$. In the region of nearly constant delay, they are approximately equal. For the chosen transfer function (11) as derived from (8):

$$t_{ph} = \frac{\theta}{\omega} = t_0\left(1 - \frac{\epsilon}{u}\right) \quad (15)$$

and

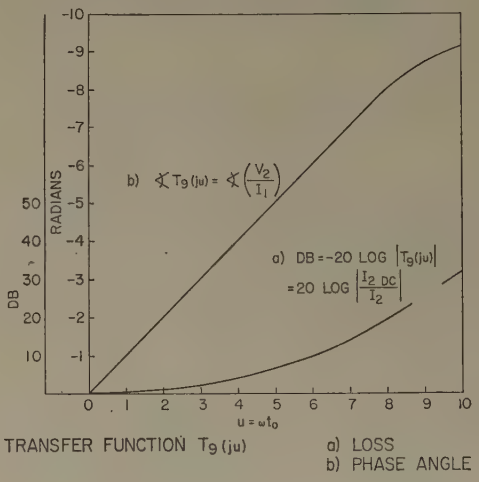
$$t_g = \frac{d\theta}{d\omega} = t_0\left(1 - \frac{d\epsilon}{du}\right) = t_0\left\{1 - \frac{1}{u^2\left[\frac{\pi}{2u}\left(J^2_{-N-1/2}(u) + J^2_{N+1/2}(u)\right)\right]}\right\}, \quad (16)$$

where t_0 is the nominal or zero-frequency delay. Expression (16) is useful for quantitative purposes, but a more vivid impression of the delay characteristic is obtained by expanding the distortion term in (16) in a McLaurin series, which certainly converges in the vicinity of $u = 0$:

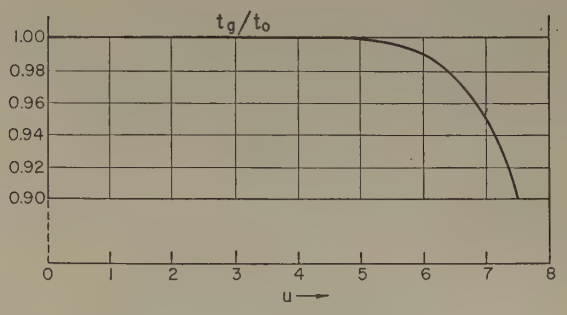
$$t_g = t_0\left[1 - \frac{u^{2N}}{a_0^2} + \frac{u^{2N+2}}{(2N-1)a_0^2} - \frac{2(N-2)}{(2N-1)^2(2N-3)a_0^2} u^{2N+4} \cdots\right]. \quad (17)$$

The absence of the terms in $u^2, u^4, \cdots, u^{2N-2}$ indicates that the first $N-1$ derivatives of t_g are zero at $u = 0$. This explains why the delay network synthesized on the basis of (8) has been characterized as "maximally flat,"[18,19] in analogy to the usage of this term in insertion loss design.[20] The delay is closely equal to its zero-frequency value t_0 up to a certain upper frequency which is a function of N, and then it declines steadily and smoothly as the frequency increases to infinity.

With equations (11) to (16) representing a complete description of the pertinent characteristics in compact form and in terms of tabulated functions, it is hardly a chore to evaluate the loss and delay distortion versus frequency for a given N and t_0, or to establish the required value of N for specified attenuation or delay distortion versus a given delay-bandwidth product. The phase-angle, group delay, and loss characteristics of $T_9(ju)$ are drawn in Fig. 1. This particular value of N has been chosen for illustration because the circuit realization and response for $T_9(ju)$ are described subsequently.



TRANSFER FUNCTION $T_9(ju)$ a) LOSS b) PHASE ANGLE



TRANSFER FUNCTION $T_9(ju)$ c) GROUP DELAY

Fig. 1

D. The Impulse Response Corresponding to $T_N(p)$.

The transfer function may be written as

$$T_N(p) = e^{\Gamma(p)} = e^{\gamma_2 p^2 + \gamma_4 p^4 + \gamma_6 p^6 + \cdots} \cdot e^{\gamma_1 p + \gamma_3 p^3 + \gamma_5 p^5 + \cdots}, \quad (18)$$

where the exponent $\Gamma(p)$ is expanded in a McLaurin series. This is permissible since $T_N(p)$ is a low-pass transfer function and analytic in the vicinity of the origin and up to the boundary set by the radius of convergence.

[18] W. E. Thomson, "Networks with maximally flat delay," *Wireless Eng.*, vol. 29, pp. 255–263; October, 1952. Corrections, *ibid.*, p. 309. (Its transfer function and $Q_N(p)$ are substantially the same.)

[19] W. H. Huggins, "Network Approximation in the Time Domain," Air Force Cambridge Res. Labs. Rep. E5048A, p. 34 *et seq.*

[20] V. D. Landon, "Cascade amplifiers with maximal flatness," *RCA Rev.*, vol. 5, pp. 347–362, 481–497; 1941.

Since, for real frequencies $p=ju$, the odd powers in the exponent of (18) determine the phase angle $(-\theta)$ [see (12)], the expression

$$t_g = \frac{d\theta}{d\omega} = t_0[-\gamma_1 + 3\gamma_3 - u^2 - 5\gamma_5 u^4 \cdots] \quad (19)$$

must agree with (17). Therefore: $\gamma_1=-1$, and $\gamma_3=\gamma_5=\gamma_7=\cdots\gamma_{2N-1}=0$. Let $f(t/t_0)$ be the impulse response corresponding to $T_N(p)$ and expand the exponential underneath the integral sign in the Laplace-transform equation, where t_0 has been inserted explicitly on the right-hand side of (20):

$$T_N(p)\cdot e^p = \int_0^\infty f(t/t_0)e^{-p(t-t_0)}dt/t_0$$

$$= \sum_{k=0}^{\infty}\left[\int_0^\infty f(t/t_0)\cdot(t-t_0)^k dt/t_0\right](-p)^k/k!. \quad (20)$$

If

$$T_N(p)\cdot e^{-\gamma_1 p} = T_N(p)\cdot e^p = e^{\gamma_2 p^2+\gamma_4 p^4+\gamma_6 p^6+\cdots}$$

$$\cdot e^{\gamma_{2N+1}p^{2N+1}+\gamma_{2N+3}p^{2N+3}+\cdots} = e^{\Gamma_c(p)}$$

is also expanded in a power series about $p=0$,

$$T_N(p)\cdot e^p = 1+\Gamma_c(p)+1/2\Gamma_c^2(p)+1/3!\Gamma_c^3(p)+\cdots,$$

and set equal to (20), it follows that the first N *odd* moments of the impulse response with respect to the mean $t=t_0$ are zero; i.e., $\int_0^\infty f(t/t_0)\cdot(t-t_0)^{2m-1}dt=0$ for $m=1, 2, \cdots N$. Consequently, the impulse response is symmetrical about $t=t_0$ to the extent that the 3rd, 5th, 7th $\cdots$ and $(2N-1)$st central moments vanish (the 1st central moment is zero by definition). Perfect symmetry is approached as $N\to\infty$, since all odd central moments tend to zero as $N\to\infty$.

An independent check is available for the above statements. If the roots of $Q_N(p)$ are introduced,

$$Q_N(p) = \prod_{r=1}^{N}(p-\rho_r),$$

then

$$\Gamma_{(p)} = \sum_{k=1}^{\infty}\gamma_k p^k = \ln T_N(p)$$

$$= \ln a_0 - \sum_{r=1}^{N}\ln(p-\rho_r). \quad (21)$$

After each term $\ln(p-\rho_r)$ is expanded in a McLaurin series, it follows that:

$$\gamma_k = \frac{1}{k}\sum_{r=1}^{N}(1/\rho_r)^k. \quad (22)$$

Burchnall[21] has shown that when $Q_N(p)$ is given by (8) and (9), then

$$\sum_{r=1}^{N}1/\rho_r = -1, \qquad \sum_{r=1}^{N}(1/\rho_r)^{2m-1} = 0,$$

$$m = 2, 3, 4, \cdots, N. \quad (23)$$

This is a welcome corroboration that the 3rd, 5th, 7th $\cdots$ and $(2N-1)$st central moments are equal to zero. Furthermore, $Q_N(p)$ is a unique solution, except for a constant multiplier, if the maximally flat delay property in accordance with (23) is postulated.

An exact description of the impulse response would require a knowledge of all the roots of $Q_N(p)$, which can be acquired only by solving for the roots of an N'th degree polynomial for each specific value of N. However, a certain amount of general information about the nature of the roots is available:[21]

(a) all the roots of $Q_N(p)$ are simple,
(b) $Q_N(p)$ has only one real zero when N is odd; if ρ_1^* is the real zero of $Q_i(p)$, then $(-1)=\rho_1^*>\rho_3^*>\rho_5^*\cdots>-\infty$,
(c) $Q_N(p)$ has no real zero when N is even,
(d) $Q_N(p)$ and $Q_{N+1}(p)$ have no root in common,
(e) all roots of $Q_N(p)$ lie outside the unit circle for $N>1$,
(f) since

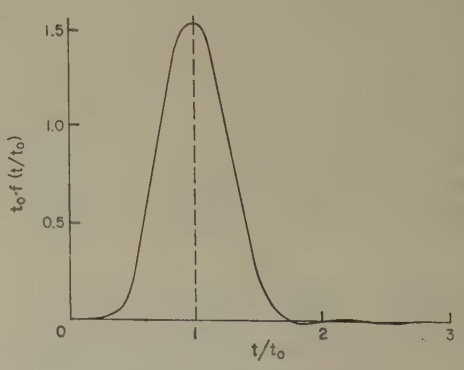
$$a_0 = a_1: \quad \sum_{r=1}^{N}1/\rho_r = -1,$$

(g) since

$$a_{N-1} = \binom{N+1}{2}: \quad \sum_{r=1}^{N}\rho_r = (-1)\cdot\binom{N+1}{2},$$

and

$$\prod_{r=1}^{N}\rho_r = (-1)^N\cdot a_0 = (-1)^N 1\cdot 3\cdot 5\cdot 7\cdot\cdots(2N-1).$$



IMPULSE RESPONSE DUE TO $T_9(p)$
$f(t)=\mathcal{L}^{-1}[T_9(p)]$

Fig. 2

The roots of $Q_N(p)$ have been tabulated in reference 18 for N from 1 to 9 and an approximation of the impulse response, which tends to a Gaussian curve for large N, has been described there.

The calculated impulse response corresponding to $T_9(p)$ is plotted in Fig. 2.

[21] J. L. Burchnall, "The Bessel polynomials," *Can. Jour. Math.*, vol. 3, pp. 62–68; 1951. E. Grosswald, "On some algebraic properties of the Bessel polynomials," *Trans. Amer. Math. Soc.*, vol. 71, pp. 197–210; September, 1951.

E. A Class of Transfer Functions for "Maximally Flat" Delay Networks

Although Part II of this paper is concerned with the realization of $T_N(p)$ exclusively, it is worthwhile to mention that a whole class of realizable transfer functions can be deduced immediately from $T_N(p)$, each member of which furnishes maximally flat delay of a certain order. This class is given by:

$$S_{M,N}(p) = \frac{Q_N(0)}{Q_M(0)} \cdot \frac{Q_M(-xp)}{Q_N(p)}, \qquad 0 \leqq M \leqq N.$$

The total zero-frequency delay equals $(1+x)t_0$, where "x" may be chosen so as to equalize the useful frequency range of $Q_M(-xp)$ and of $Q_N(p)$. $S_{M,N}(p)$ appears to have a smaller total delay-bandwidth product in general than $T_{M+N}(p)$, but its rate of transmission loss is lower in comparison. As M approaches N, "x" approaches "1" and the loss-bandwidth continues to increase. Finally, $M=N$ corresponds to an all-pass network with a zero-frequency delay of $2t_0$. The only member of this class which is a minimum-phase transfer function is $S_{0,N}(p)=T_N(p)$. All others possess zeros in the right half of the p-plane.

The loss and delay characteristics of $S_{M,N}(p)$ can be obtained by combining

$$L = 20 \log (A_M/A_N),$$

$$t_{ph} = t_0[1 + x - (\epsilon_M + \epsilon_N)],$$

and

$$t_g = t_0\left[1 + x - \left(\frac{d\epsilon_M}{du} + \frac{d\epsilon_N}{du}\right)\right]$$

with (13), (14) and (16); in the evaluation of A_M and ϵ_M, "xu" should replace "u" as the argument. Strictly speaking, the property of maximally flat delay is tied to M rather than N in case of $S_{M,N}(p)$, except that the smallness of "x" for small M's minimizes the deviations from zero of the terms in u^{2M}, u^{2M+2}, $\cdots$, u^{2N-2}, occurring in the expansion of t_g. On the other hand, the deviations are emphasized and the loss is reduced further, or even turned into gain over a certain extent of "u," by making "x" larger than in accordance with the above recommendation. The introduction of the nonminimum-phase transfer functions $S_{M,N}(p)$, involving the two-fold choice of M and "x," offers more variety in amplitude vs. phase characteristics and even vs. odd moments of the impulse response than the simpler functions $T_N(p)$.

If the realization of $S_{M,N}(p)$ is to be cast in the form of a reactive or quasi-reactive (see Part II) network terminated in a resistance, it is necessary first to multiply numerator and denominator by a Hurwitz polynomial which makes the over-all numerator an even function of the frequency variable used in the lc development. If this variable is "p," then the multiplier is $Q_M(xp)$ and the class of modified transfer functions is given by:

$$\bar{S}_{M,N}(p) = \frac{Q_N(0)}{Q_M(0)} \cdot \frac{Q_M(-xp)Q_M(xp)}{Q_N(p)Q_M(xp)}, \qquad 1 \leqq M \leqq N.$$

Part II

The Ladder Network Realization of $T_N(p)$

In Part I a rather convincing case has been made with regard to the suitability of $T_N(p)$ as the transfer function of a low-pass lumped-parameter delay network. Also, fairly simple expressions have been set up in terms of tabulated functions to guide the choice of N in meeting given requirements regarding delay, bandwidth, and distortion. It is now necessary to find a suitable configuration of specific resistances, inductances, and capacitances, which actually possesses $T_N(p)$ as its transfer function. It is particularly desired that the network be unbalanced and that coils with only low Q's be required in its construction. Since the source of the signal is likely to be a vacuum tube, i.e., a high-impedance (constant-current) generator, the emphasis will be placed on realizing $T_N(p)$ as a transfer impedance, so that $T_N(p)$ will equal the ratio of output voltage to input current of the resulting network. The method of designing a network so that $T_N(p)$ is represented by the ratio of output voltage to input voltage is quite analogous and is useful when dealing with a low-impedance (constant-voltage) generator.

A. Network Synthesis on a Quasi-Reactive Basis.

The choice of a constant as the numerator in the transfer function (8) opens the way to an uncomplicated method of synthesis, as well as to a very attractive network configuration. The key to the dominating position occupied by a constant in the numerator is the fact that it makes the network realizable as a ladder structure, and that the numerator retains its evenness in the frequency variable under the transformation $p=s-d$,[22] where "s" is an auxiliary frequency variable and "d" a constant. The combination of these two properties amounts to a sufficient condition in making it possible to develop the network on a quasi-reactive basis in the auxiliary "s" domain by the straightforward Cauer process,[23] but to arrive at a final network that has a uniform loss component attached to each reactance. The loss components may be proportioned so as to reduce the required Q's of the coils to rather low values at the expense of a fixed transmission loss. It goes without saying that the successful restriction of the actual synthesis process to the 2-element lc case is rather attractive as compared to a general rlc method of synthesis.

The transfer impedance of a two-terminal-pair, finite, linear, bilateral network, terminated by a normalized

[22] S. Darlington, "Synthesis of reactance 4-poles," *Jour. Math. Phys.*, vol. 18, pp. 257–333; September, 1939. (See part IV, "Dissipative Reactance Networks.") Also, H. W. Bode, *op. cit.*, pp. 216–223.

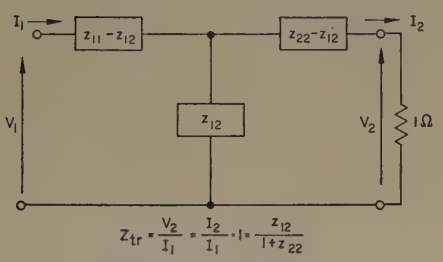
[23] A. E. Guillemin, "Modern Methods of Network Synthesis," *Advances in Electronics*, Academic Press, New York, N. Y., vol. III, p. 286, *et seq.*

load resistance of 1 ohm, is easily derived from the equivalent T-section, in Fig. 3, as:

$$Z_{tr} = \frac{V_2}{I_1} = \frac{z_{12}}{1 + z_{22}} \cdot \tag{24}$$

Let "d" be chosen such that the shift of the real-frequency axis to the left, implied in the translation $p = s - d$, does not carry it into or past any of the roots of $Q_N(p)$, and let the transformation $p = s - d$ be carried out in (8), obtaining:

$$T_N^*(s) = \frac{a_0}{Q_N^*(s)} = \frac{a_0}{G_N^*(s) + H_N^*(s)} = \frac{a_0}{\sum_{r=0}^{N} a_r^* s^r} \cdot \tag{25}$$



EQUIVALENT-T NETWORK

Fig. 3

Here $G_N^*(s)$ and $H_N^*(s)$ represent the even and odd parts, respectively, of $Q_N^*(s)$. Since the numerator is even in "s," (25) may be realized within a real multiplier as the transfer impedance of low-pass, reactive network as follows. Let the constant in the numerator be changed to a_0^*, in order to make the dc value ($s = 0$) of (25) equal to "1," as required for the transfer impedance of a low-pass reactive network when ideal transformers are excluded and the termination is a 1-ohm resistor. After re-arranging the terms so that

$$Z_{tr}^* = \frac{\dfrac{a_0^*}{H_N^*(s)}}{1 + \dfrac{G_N^*(s)}{H_N^*(s)}}, \tag{26}$$

the following identifications are prominent in view of (24):

$$z_{12}^* = \frac{a_0^*}{H_N^*(s)} \tag{27}$$

$$z_{22}^* = \frac{G_N^*(s)}{H_N^*(s)} \cdot \tag{28}$$

Surely, $z_{22}^* = G_N^*(s)/H_N^*(s)$ is a driving-point reactance function inasmuch as $Q_N^*(s)$ is a Hurwitz polynomial by hypothesis (see Part I, section A). Furthermore, $z_{12}^* = a_0^*/H_N^*(s)$ is an odd function of "s," its poles are also poles of $z_{22}^* = G_N^*(s)/H_N^*(s)$, and the numerator is of lower degree than the denominator. Therefore, the Z_{tr}^* of (26) is realizable by a reactive network.

A-1. *The Ladder Development:* Synthesis by means of Cauer's straightforward development of a reactive ladder network[23] is particularly applicable, since the structure of the numerator is so simple. This method does not require finding the roots of the N'th degree equation $Q_N^*(s) = 0$ and it leads to a network which requires only N reactances for an N'th degree transfer function (8), which is the smallest possible number.[24] The principal step required by this method is the expansion of the single reactance function $\psi(s)$ in a continued fraction in "s":

$$\psi(s) = \frac{s^N + a_{N-2}^* s^{N-2} + a_{N-4}^* s^{N-4} + \cdots}{a_{N-1}^* s^{N-1} + a_{N-3}^* s^{N-3} + a_{N-5}^* s^{N-5} + \cdots}$$

$$= \alpha_N^* s + \cfrac{1}{\alpha_{N-1}^* s + \cfrac{1}{\ddots + \cfrac{1}{\alpha_1^* s}}} \cdot \tag{29}$$

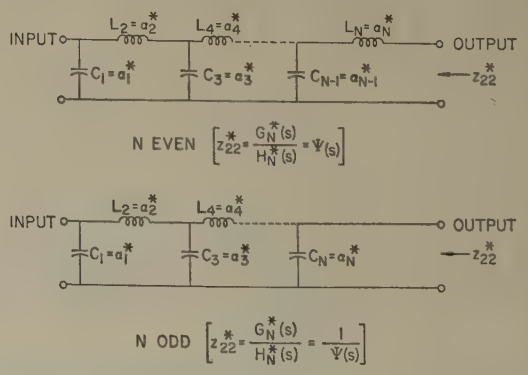
The procedure for obtaining the coefficients α_j^* ($j = 1, 2, \cdots, N$) of the continued fraction, which must all be non-zero and positive, is explained in reference 11. If zero or negative coefficients should appear, it would indicate that "d" has been chosen too large or, much more likely, that not enough significant figures have been retained in calculating the a_r's and in the subsequent computations, which may happen when N is quite large.

It is quite apparent that when N is even

$$\psi(s) = \frac{G_N^*(s)}{H_N^*(s)} = z_{22}^* \tag{30}$$

but that when N is odd

$$\psi(s) = \frac{H_N^*(s)}{G_N^*(s)} = \frac{1}{z_{22}^*} \cdot \tag{31}$$



LADDER DEVELOPMENT OF z_{22}^*

Fig. 4

The ladder network for z_{22}^* may be readily diagrammed for even or odd N's in terms of the coefficients α_j^*, as is shown in Fig. 4.[11] The computations of the component

[24] A. E. Guillemin, "Communication Networks," John Wiley and Sons, Inc., New York, N. Y., vol. II, p. 203.

values may be checked by

$$\sum_k \alpha^*_{2k-1} = \frac{a_1^*}{a_0^*} \quad \text{and} \quad \prod_{k=1}^{N} \alpha_k^* = \frac{1}{a_0^*}.$$

A degree-reducing, common factor of $G_N^*(s)$ and $H_N^*(s)$ would have to be an even polynomial in "s," which would possess roots on the axis $Re[s]=0$ or in the right half of the s-plane. But such a factor must be ruled out, since it would also be contained in $Q_N^*(s)=G_N^*(s)+H_N^*(s)$, and thereby violate the condition that $Q_N^*(s)$ must be a Hurwitz polynomial. The ladder development of Fig. 4 actually corresponds to (28) and not just to some submultiple thereof.

If a pair of input terminals is put across C_1, for N odd or even, then z_{12}^* has no finite zeros of transmission, since none of the network arms is resonant at a finite, non-zero frequency; furthermore, z_{22}^* can have at most a pole at $s=\infty$ in addition to the poles of z_{12}^*, which occurs when N is even and there is a coil in series with the output terminals. Therefore, $z_{12}^*=a_0^*/H_N^*(s)$ as required by (27). All that remains to be done is to connect a 1-ohm load resistance across the z_{22}^*-terminals (output terminals), in order to complete the network realization of Z_{tr}^* of (26).

Nothing has been said so far about z_{11}^* because it does not affect the transfer impedance at all. It will be shown in section *C.* that the input impedance of the terminated network is finite in $Re[s]\geqq 0$. This implies that the residue condition applicable to lc networks[25] is fulfilled with the equal sign, and z_{11}^* may be determined on this basis without resorting to the circuit diagram.

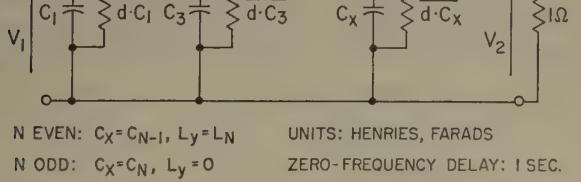


Fig. 5

A-2. Final Network. To remove the effect of the initial frequency transformation $p=s-d$, the network arms $s\cdot L_{2k}$ and $s\cdot C_{2k-1}$ must be replaced by $L_{2k}(p+d)$ and $C_{2k-1}(p+d)$, respectively; i.e., a resistor $d\cdot L_{2k}$ must be placed in series with the inductance L_{2k} and a resistor $1/d\cdot C_{2k-1}$ must be shunted across the capacitor C_{2k-1}, where $k=1, 2, 3, \cdots, N/2$ (for N even) or $(N\pm1)/2$ for N odd). By (8), (25), and (26), the resulting network of Fig. 5 has the transfer impedance

$$Z_{tr} = \frac{a_0^*}{a_0} T_N(p), \tag{32}$$

which equals $T_N(p)$ except for a constant multiplier. The price that has to be paid for the insertion of the resistive components is the fixed loss corresponding to $k=a_0^*/a_0$, which may be calculated from $a_0^*=Q_N(p)\big|_{p=(-d)}$. For small values of "$d$" ($d\ll 1$), the loss is somewhat less than 20 log $1/1-d$ db. Finally, the normalizations are removed by multiplying all L_{2k}'s by $t_0\cdot R_L$, all C_{2k-1}'s by t_0/R_L, and all R's by R_L, if a nominal delay t_0 (in seconds) and a load impedance level R_L (in ohms) are desired. The new impedance level introduces a factor R_L in (32), i.e., $Z_{tr}=(a_0^*R_L/a_0)T_N(p)$. If Q has its conventional meaning, then $Q=1/d$ at $\omega_0=1/t_0$ for all network arms and varies linearly with frequency.

B. Illustrative Example: The Realization of $T_9(p)$

The effect of increasing an even N by "1" is to add a capacitor across the output terminals without changing the number of coils in the network. One will usually, therefore, prefer an odd N in order to maximize the delay-bandwidth product for a given number of coils, usually bulkier and more expensive than capacitors. Let $N=9$, which possesses a delay-bandwidth product suitable for an application the writer had to deal with, and let $t_0=1.25$ milliseconds, $R_L=4{,}000$ ohms. The loss (with respect to dc) and delay properties of $T_9(ju)$ are shown in Fig. 1; its "u" scale may be converted to a direct frequency scale corresponding to $t_0=1.25$ milliseconds by multiplying each value of "u" by 128 cps.

First, the a_r's of (9) have to be calculated for $N=9$. This may be done with the help of factorial tables,[26] or by a direct method after writing out the terms and cancelling common factors in numerator and denominator. The polynomial $Q_N(p)$ may also be obtained by recursion [see (6)]:

$$Q_k(p) = (2k-1)Q_{k-1}(p) + p^2Q_{k-2}(p), \tag{33}$$

with $Q_{-1}=1/p$, $Q_0=1$; but this method is not efficient for large N's. One obtains for $Q_9(p)$:

$$\begin{aligned} Q_9(p) = \sum_{r=0}^{9} a_r p^r = {} & p^9 + 45p^8 + 990p^7 + 13{,}860p^6 \\ & + 135{,}135p^5 + 945{,}945p^4 + 4{,}729{,}725p^3 \\ & + 16{,}216{,}200p^2 + 34{,}459{,}425p \\ & + 34{,}459{,}425. \end{aligned} \tag{34}$$

Secondly, a value of "d" is chosen on the basis of the coils to be used or of the fixed loss that can be tolerated. The upper limit on "d," imposed by the condition of physical realizability, is not likely to be a hindrance in practice, since it is 1.5 for $N=2$ and higher for larger N's; e.g., $d_{max}\to 2.9$ for $N=9$. The loss, of course, increases as "d" grows larger. A preliminary analysis indicates that for $t_0=1.25$ milliseconds and $R_L=4{,}000$ ohms, which were required values for the particular application, the maximum required inductance will be about

[25] A. E. Guillemin, "Modern Methods of Network Synthesis," *Advances in Electronics*, Academic Press, N. Y., vol. III, p. 274; 1951. W. Cauer, "Theorie der linearen Wechselstromschaltungen," Becker und Erler, Leipzig, Ger., vol. I, p. 373; 1941.

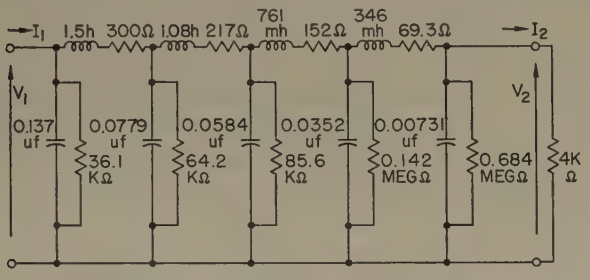
[26] NBS, Applied Math. Series 16, "Tables of $n!$ and $\Gamma_{(n+1/2)}$ for the First Thousand Values of n," U. S. Dept. of Commerce, 1951.

1.5 henry. If a toroidal molybdenum-permalloy dust core of 0.8 inch outside diameter is decided upon, the minimum Q among the coils equals 4 at $f_0=128$ cps ($f_0=1{,}000/2\pi\cdot 1.25=128$ cps). Therefore, the value of "d" may be fixed at 0.25, which corresponds to a fixed loss of 2.2 db for $N=9$.

Substituting $p=s-0.25$ in (34), the a_r*'s may be obtained by means of synthetic division. The large number of significant figures is due to machine calculation and not indicative of required accuracy. Expanding the ratio of the odd to the even part

$$\begin{aligned}\psi(s) = {} & (s^9 + 902.25s^7 + 115{,}605.492s^5 \\ & + 3{,}864{,}041.03s^3 + 27{,}181{,}586.9s) \\ & \div (42.75s^8 + 12{,}204.9375s^6 + 789{,}490.775s^4 \\ & + 13{,}003{,}312.9s^2 + 26{,}787{,}745.75)\end{aligned}$$

in a continued fraction, introducing the loss components, and removing the normalizations, the network of



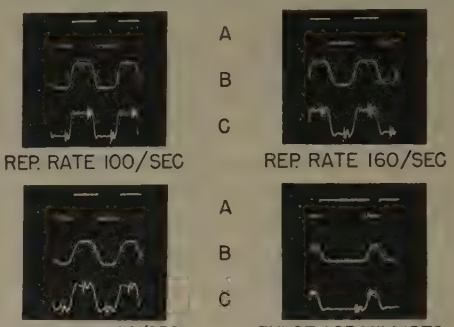
MAXIMALLY FLAT DELAY NETWORK FOR $T_{9(p)}$
$R_L=4K\Omega$, $t_0=1.25$ MILLISEC., $Q=4$ AT 128 cps

$$\frac{I_2}{I_1}\rightarrow 0.777\epsilon^{-1.25\cdot 10^{-3}\omega}$$

Fig. 6

Fig. 6 is obtained. It may be worth noticing that the spread in the values of the elements is not too large and that a generator impedance higher than 29,200 ohms can be accommodated without effecting the exact transfer characteristic. A change in the load impedance would, of course, effect this latter value in the same proportion.

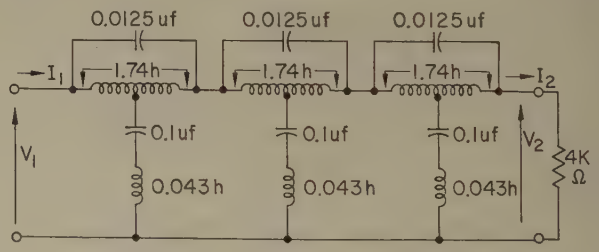
The ladder configuration makes it possible to dispense with stringent tolerances for the network components, as required for the conventional bridged-T image-parameter delay networks. Even so, the measured amplitude and phase response are indistinguishable from the calculated curves of Fig. 1. The time-domain behavior can be judged by Fig. 2 and the oscillograms of Fig. 7. For each square-wave input of Fig. 7A, the response of the maximally flat delay network of Fig. 6 is shown in Fig. 7B, and that of a bridged-T delay network[27] of comparable delay and linear phase-shift bandwidth is shown in Fig. 7C. The particular bridged-T delay network of Fig. 8 was chosen for comparison primarily because of its availability and not because it represents the ultimate in image-parameter design. In any case, the superiority of the square-wave response obtained from the maximally flat delay network is rather pronounced. The more than 12:1 reduction in weight and volume achieved by means of the present design may be of considerable interest in view of today's prevalent need for miniaturization: the total weight of the coils is only 2 ounces for the maximally flat delay network of Fig. 6, but it exceeds 1½ pounds for the bridged-T delay network of Fig. 8 (coil Q's are about 30 at 128 cps), with a similar ratio for the relative volumes.



SQUARE-WAVE RESPONSE OF 1.25 MILLISEC. DELAY NETWORKS: B) MAXIMALLY-FLAT; C) BRIDGED-T.

Fig. 7

On account of its high-grade pulse response, the maximally flat delay network should qualify for applications that require a substantially nonoscillatory transient response, without the time delay being of primary importance.



BRIDGED-T DELAY NETWORK
(OVER-ALL DELAY 1.25 MILLISEC.)

$$\frac{V_2}{V_1}=\frac{I_2}{I_1}\longrightarrow \epsilon^{-j1.25\cdot 10^{-3}\omega}$$

(ACTUAL COIL Q's ABOUT 30 AT 128 cps)

Fig. 8

A comparison may also be made with a 4-section constant-k filter delay network, which has the same general configuration as the network of Fig. 6. Its t_g would differ from t_0 by one per cent at somewhat less than $u=1.3$, assuming even infinite Q's for the coils. The same error occurs for the maximally flat delay network of Fig. 6 not until $u=6.08$, and even its loss is not down by 3 db from the dc value until "u" reaches 3.38. The only debit that can be charged against the maximally flat delay network, in competing with the constant-k filter, involves cascade-connection, which is discussed further in section C. Incidentally, the coils and capacitors of the constant-k filter are much larger for the same R_L

[27] Designed by J. E. Taber of the Hughes Research and Development Laboratories (see Figure 8).

and t_0, being 5 henry for the coil and 0.31 μf for the capacitors of each section. These comparisons are a measure of the improvements achieved by transfer-function synthesis as compared to classical design of a network with a similar configuration.

C. Cascading of Delay Networks of the Type of Fig. 6

There are applications that require several equal or unequal increments of delay rather than a single fixed value. When the over-all fixed transmission loss is not excessive, the question arises how maximally flat delay networks, designed on the transfer impedance basis, can be cascaded without having to insert isolating amplifiers. All prerequisites for cascade-connection are fulfilled if the input impedance Z_{in} is equal to a constant resistance, preferably R_L, independent of frequency. This is not the case for the synthesis procedure of section B., which controls z_{12} and z_{22} but does not shape z_{11}, at the expense of additional elements, in order to maintain a constant input-impedance level. This method of synthesis does, however, produce an input impedance that is finite along the whole real-frequency axis and thereby paves the way for transforming the network into a "constant-resistance" filter.

If the transmission matrix for the network of Fig. 5 is given by

$$\begin{pmatrix} V_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} \tilde{A}, \tilde{B} \\ \tilde{C}, \tilde{D} \end{pmatrix} \cdot \begin{pmatrix} V_2 \\ I_2 \end{pmatrix}, \tag{35}$$

then

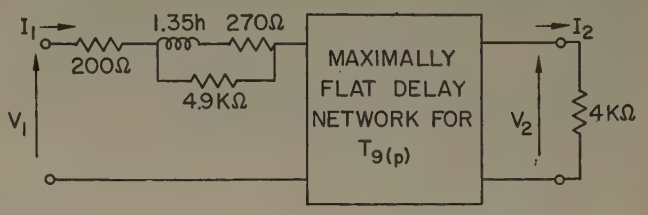
$$Z_{in} = \frac{\tilde{A} + B}{\tilde{C} + \tilde{D}}. \tag{36}$$

For a ladder network, the elements of the transmission matrix are entire rational functions of the impedances of the series arms and of the admittances of the shunt arms. Since these are $L_{2k} \cdot (p+d)$ and $C_{2k-1} \cdot (p+d)$ in Fig. 5, $\tilde{A}$, $\tilde{B}$, $\tilde{C}$, and $\tilde{D}$ are all polynomials in "p." Inasmuch as

$$Z_{tr} = \frac{V_2}{I_1} = \frac{1}{\tilde{C} + \tilde{D}} \tag{37}$$

in the present notation, the denominator of (36) cannot differ from the Hurwitz polynomial $Q_N(p)$ by more than a constant. It follows conclusively that Z_{in} is finite at least everywhere in $Re[s] \geqq 0$ and, in particular, along the whole real-frequency axis $p = j\omega$, including $\omega = \infty$ because of the presence of C_1. This being the case, a driving-point impedance Z^c_{in} can be developed such that $Z_{in} + Z^c_{in}$ equals a constant for all real frequencies. A network arm connected in series with the input terminals does not alter the transfer impedance Z_{tr} of a network. Therefore, Z^c_{in} may be connected in such a manner, to establish a fixed input-impedance level and enable cascading of the desired number of these networks, designed for equal or different amounts of delay.

While this method of input-impedance compensation is exact, it would appear to be rather wasteful in element for practical purposes. The number of elements needed to construct Z^c_{in} would equal the number used in the original network of Fig. 6. No general approximate solution has been attempted, but a rather serviceable approximate compensating impedance has been worked out readily for the network realization of $T_9(p)$ (Fig. 6). The added series arm, shown in Fig. 9, and consisting only of a few elements, is sufficient for good compensation over the major portion of the pass-band. Several of the networks of Fig. 9 have been operated in cascade with complete success, undistorted and properly delayed outputs having been taken from each junction point of two of these networks and at the termination.



MAXIMALLY FLAT DELAY NETWORK FOR CASCADE—CONNECTION

$R_L = 4K\Omega$, $t_0 = 1.25$ MILLISEC., $Q = 4$ AT 128 cps

$$\frac{V_2}{V_1} \cong \frac{I_2}{I_1} \rightarrow 0.777 e^{-j1.25 \cdot 10^{-3}\omega}$$

Fig. 9

It should be realized that the cascading of maximally flat delay networks increases the rate of transmission loss, since the loss curve per network is bell-shaped. More precisely, a total of N reactances distributed equally among "n" networks, each of degree N/n and contributing a delay t_0/n, has a 3 db bandwidth of

$$u_{3db} \cong \sqrt{(2N - n) \ln 2}, \tag{38}$$

but a network designed directly on the N'th degree basis has a 3 db bandwidth of

$$u_{3db} \cong \sqrt{(2N - 1) \ln 2}, \tag{39}$$

which is $\sqrt{(2N-1)/(2N-n)}$ times larger than (38). It stands to reason that cascading normally would only be used when the additional tap-points are required. Otherwise, the network should be synthesized on the basis of the full N, since in this manner a greater useful delay-bandwidth product is obtained at the expense of more lengthy calculations.

Acknowledgment

The author is indebted to Dr. Louis Weinberg for stimulating discussions on network synthesis and to Mr. F. G. Rasmussen for making laboratory tests and taking the oscillograms, both of whom are with the Research and Development Laboratories, Hughes Aircraft Company. He is further indebted to Mr. W. E. Thomson for indicating in a private communication the relationship of Lommel polynomials to reference 18.

Gain-Bandwidth Limitations on Equalizers and Matching Networks*

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Summary—If an equalizer amplitude-response curve is specified, it will be shown that the minimum flat loss obtainable with physical networks is determined. This flat loss, or scale factor on the response curve, is a function of the equalizer output-terminating impedance, which is arbitrary but prescribed, and the specified tolerance on input mismatch.

If the output impedance is purely reactive, the limitations on maximum voltage transfer are obtained from a consideration of the open circuit impedance parameters of the system. If power or voltage transfer to a load with finite real part is to be optimized, the scattering parameters of the system are used to determine the limits of performance.

Examples will be given comparing the performance of matched and lossless equalizers. In many practical cases the latter do not have substantially higher gain than the matched equalizer.

Definition of Equalizer Problem

THE EQUALIZER problem considered here concerns the transfer of voltage, current, or power from a prescribed generator with resistive internal impedance to a load whose impedance is a given function of frequency. It is presumed that a real frequency function is specified which defines the shape of transfer gain characteristic desired, and it is required to find a passive linear reciprocal equalizer network (a two terminal pair transducer, or two-port) which, when placed between generator and load, produces the specified gain shape and does so with maximum scale factor, i.e., minimum flat loss. A gain characteristic that ideally is constant over a finite frequency band and zero elsewhere will be of major interest, and many of the results given may therefore be regarded as generalizations of the concept of maximum "gain-bandwidth product."[1]

An additional specification is the tolerance on input mismatch. It is only because the equalizer networks investigated here are not limited to the lossless case that this specification can be set independently of the others. In most of the examples given below the extremes of a matched input dissipative equalizer (zero mismatch) and a completely lossless equalizer (but not matched) will be compared.

Various aspects of the equalizer problem have been previously considered. Bode[1] discusses the limitations on "gain-bandwidth" product imposed by a load with shunt capacitance when a lossless equalizer is used, and also gives some consideration to matched equalizers for voltage transfer to a reactive load.[2] Fano[3] has treated the problem of optimum match of an arbitrary load with a lossless network, and La Rosa and Carlin[4–6] have examined this problem when the lossless restriction on the matching network is removed. Norde[7] has treated matched minimum phase voltage equalizers for reactive loads. Other work on special aspects of "gain-bandwidth" product is too extensive to be given here. Wheeler[8] and Hansen[9] are typical references.

This paper considers the general approach to any equalization problem, and stems directly from the references cited above. The results presented on optimum voltage transfer to an arbitrary load (including the purely reactive load case) have not been given elsewhere.

General Approach to Equalization of an Arbitrary Load

There are two basic restrictions that govern the design of an equalizer network. One of these is the general requirement of physical realizability on the over-all network which includes both equalizer and prescribed load. The other is the total set of constraints specifically imposed by the load, and this should be entirely independent of the equalizer network. If these constraints are satisfied, then, when the over-all network is synthesized and the given load removed, the remaining circuit (the equalizer alone) is physically realizable. The form in which these restrictions are stated must be such as to involve explicitly (and preferably in a simple way) the transfer gain function, the scale factor of which is to be maximized. The process of finding the optimum equalization is then to adjust the gain function within these general restrictions until the limits of physical realizability are attained.

The constraints that apply to the equalization of a load containing dissipation (the purely reactive load case is considered later) are most readily obtained by

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[1] H. W. Bode, "Network Analysis and Feedback Amplifier Design," D. Van Nostrand Book Co., New York, N. Y.; 1945.

[2] H. W. Bode, U. S. Patent 2,337,965, 1942.

[3] R. M. Fano, "Theoretical limitations on the broadband matching of arbitrary impedances," *Jour. Frank. Inst.*, vol. 249, pp. 57–83; January, 1950; vol. 249, pp. 139–154; February, 1950.

[4] R. La Rosa, "Broadband Dissipative Matching Networks," D.E.E. Thesis, Polytechnic Institute of Brooklyn; May, 1953.

[5] R. La Rosa and H. J. Carlin, "A General Theory of Wideband Matching," Microwave Research Institute, Polytechnic Institute of Brooklyn, Report R-308-53, PIB-247; February, 1953. (To be published in *Jour. Math. & Physics.*)

[6] H. J. Carlin and R. La Rosa, "Broadband reflectionless matching," *Proc. Sym. Modern Network Syn.*, vol. 1, pp. 161–178, Polytechnic Institute of Brooklyn; April, 1952.

[7] L. Norde, "Broadband Matching with Maximum Gain and Reactive Load," M.E.E. Thesis, Polytechnic Institute of Brooklyn; June, 1953.

[8] H. A. Wheeler, "The maximum speed of amplification in a wideband amplifier," Wheeler Monographs, vol. 1, 1953. (Original publication 1949).

[9] W. W. Hansen, "On maximum gain bandwidth product in amplifiers." *Jour. Appl. Phys.*, vol. 16; September, 1945.

representing the prescribed load over the infinite frequency spectrum as a purely reactive two-port with fixed elements terminated in a unit resistor. (Hereafter the generator resistance will be presumed normalized to unity). This representation is always possible and the single resistor is sufficient to account for all the power dissipated in the load.[10,11] The "over-all network" is now defined as the equalizer plus the reactance two-port portion of the load.

In order that the "over-all network" be physically realizable, it must have an array of scattering coefficiencts[12] $S_{11}(p)$, $S_{22}(p)$, $S_{12}(p)$ which form the matrix of a positive definite or semi-definite Hermitian form for $Re\ p \geqq 0$.[5,6,12,13] The algebraic expression of this requirement gives the general set of realizability constraints previously referred to.

The specific load constraints are obtained from the fact that at certain real and complex frequencies no power can be transferred to the load no matter what equalizer network is used. These frequencies are the points on $j\omega$ and in the right hand half of the p plane at which the reactive two-port portion of the load has zeros of transmission. At these frequencies the transmission factor of the over-all network $S_{12}(p)$ must generally have a zero of transmission of order $2n$ if the load zero is of order n. Further, the reflection factor looking in at the back end of the "over-all network," i.e., $S_{22}(p)$, and generally its first $2n-1$ derivatives are completely determined by the reactance two-port portion of the load. These properties follow from a consideration of the scattering equations for the cascade connection of a pair of two-ports,[3] and constitute the "load constraints" referred to earlier. It must be emphasized that load constraints are independent of the equalizer.

The load constraints amount to the statement that essentially the first $2n$ Taylor coefficients of the back end reflection factor of both load and over-all network are equal in the series expansion about a load zero of transmission.

These requirements may be expressed in terms of the Cauchy formulas for the Taylor coefficients, and as a final result one obtains integral formulas for the logarithm of the amplitude of the back end reflection factor. If the over-all network is specified so that it satisfies the general realizability requirements and in addition meets the limits on $\ln|1/S_{22}(j\omega)|$ imposed by the integral formulas, then the prescribed load can always be separated from the over-all network leaving a physically realizable equalizer two-port. A statement of these realizability conditions in the form of a theorem essentially as given by La Rosa[4,5] is as follows:

Theorem 1:

The necessary and sufficient conditions that a scattering matrix $[S(p)]\,(p=\sigma+j\omega)$ represent an over-all network composed of an equalizer in tandem with a prescribed lossless two-port (the reactance two-port portion of the prescribed load) is:

(a) Matrix $[S]$ should be realizable, i.e., $I-[S(j\omega)]^{*T}[S(j\omega)]$ must be the matrix of a positive definite or semi-definite Hermitian form, with $[S]$ symmetric and its elements rational functions of p with real coefficients, and no right half plane poles.[13]

(b) Right hand and boundary zeros of transmission of the load must appear in the transmission factor $S_{12}(p)$ of the over-all network with at least the same multiplicity.

(c)[14] A set of integral restrictions on $\ln|1/S_{22}(j\omega)|$ of the form

$$\int_0^\infty f_i(\omega)\ln\left|\frac{1}{S_{22}(j\omega)}\right|d\omega = K_i \qquad (1)$$

must be simultaneously satisfied at all the zeros of transmission of the load. Each nth order zero contributes N_i integral equations with

$$N_i = \begin{cases} n & \text{for a zero at zero or infinity} \\ 2n & \text{for a zero on } j\omega \\ 2n-n_0 & \text{for a right hand zero on the real axis} \\ 4n-2n_0 & \text{for a conjugate pair of zeros in the right half plane.} \end{cases}$$

n_0 is the order of any right half plane zero of load transmission coincident with a zero of back end load reflection factor, i.e., of the reactance two-port portion of the load.

Theorem 1 can be applied directly and simply to a variety of equalization problems involving a load containing dissipative elements. Special consideration will be required for problems involving a purely reactive load.

Power Transfer Equalization

The application of the theorem given in the above discussion requires a determination of the relationships between the transfer function which is to be optimized and the reflection factor amplitude of the over-all network $|S_{22}(j\omega)|$. The integral equations can then be used to determine "gain-bandwidth" type of restrictions on the equalizer. Part (a) of the theorem contains the necessary information for relating the reflection factor function $S_{22}(j\omega)$ to the equalization response of the over-all network. In the case of power transfer from a generator with unit internal impedance to a load (represented in Darlington form), the insertion power gain at real frequencies of the over-all network normalized to the available generator power is

$$\frac{P_L}{P_0} = |S_{12}(j\omega)|^2. \qquad (2)$$

[10] S. Darlington, "Synthesis of reactance 4-poles," *Jour. Math. & Physics*, vol. 18, pp. 257–353; September, 1939.

[11] H. J. Carlin and R. La Rosa, "On the synthesis of reactance 4-poles," *Jour. Appl. Phys.*, vol. 24; October, 1953.

[12] A. E. Laemmel, "Scattering matrix formulation of microwave networks," *Proc. Sym. Modern Networks Syn.*, vol. 1, Polytechnic Institute of Brooklyn; April, 1952.

[13] V. Belevitch, "Synthese des Reseaux Electriques Passifs a n-Paires de Bornes de Matrice de Repartition Predeterminee." Annales des Telecommunications, vol. 6, p. 302; November, 1951.

[14] The weighting functions $f_i(\omega)$ are tabulated by Fano.[3] The K_i are related to the Taylor coefficients of the load at the zeros of transmission and are also tabulated in the same reference.

P_L is the power delivered to the load, P_0 is the available generator power ($|V_g|^2/4$, where V_g is the generator open-circuit voltage) and $|S_{12}(j\omega)|$ is the amplitude of the voltage transmission coefficient of the over-all network. Since this power transfer function is an element of the scattering matrix, it is directly related to $|S_{22}(j\omega)|$ by the general realizability constraints of part (a) of the theorem. La Rosa[5,6] has shown that this portion of the theorem leads to the following necessary requirement on $|S_{12}(j\omega)|$ for an equalizer that maximizes the scale factor of power function when the shape is specified.

$$|S_{12}(j\omega)|^2 = (1 - |S_{22}(j\omega)|)(1 + |S_{11}(j\omega)|)|S_{22}(j\omega)| \geqq |(S_{11}(j\omega)|. \quad (3)$$

In (3), $|S_{11}(j\omega)|$ is a specified input reflection factor amplitude function which sets the tolerance on input mismatch. In the special case that the equalizer is lossless, $|S_{11}(j\omega)| = |S_{22}(j\omega)|$ and (3) becomes

$$|S_{12}(j\omega)|^2 = 1 - |S_{22}(j\omega)|^2 \text{ (lossless equalizer).} \quad (4)$$

In another special case where the input mismatch is zero, i.e., $S_{11}(p) = 0$, (3) reduces to

$$|S_{12}(j\omega)|^2 = 1 - |S_{22}(j\omega)| \text{ (matched equalizer).} \quad (5)$$

Eqs. (3), (4), (5) give the desired relations between $|S_{22}(j\omega)|$ and the power transfer function $|S_{12}(j\omega)|^2$. The integral equations, (1), may therefore be expressed in terms of the power transfer function and solved to obtain the maximum scale factor. The details of this procedure as well as examples are given in references.[4-6] The solution of the equations always gives a unique maximum-gain scale factor for a prescribed shape of power transfer function and this cannot be exceeded by any physical equalizer. In the case where the equalizer is to produce a flat pass band with zero gain outside this band, the solution for the scale factor is particularly simple and is found directly in terms of a minimum constant value of $|S_{22}(j\omega)| = |S_{22}|$ over the prescribed band with $|S_{22}(j\omega)| = 1$ elsewhere. In this case it is interesting to compare the optimum lossless and matched equalizers using (4) and (5)

$$10 \log \frac{|S_{12}|^2 \text{ (lossless)}}{|S_{12}|^2 \text{ (matched)}} = 10 \log (1 + |S_{22}|) \leqq 3db \quad (6)$$

since $|S_{22}| \leqq 1$. In any practical design of a flat power equalizer $|S_{22}|$ is considerably less than 1, so that over a specified band the gain of an optimum matched design is much closer to the gain of an optimum lossless equalizer than the outside limit of 3 db given by (6).

Voltage Transfer Equalization of General Dissipative Load

A. Integral Constraints for Voltage Transfer

Theorem 1, given above, may be applied to the problem of voltage equalization, provided the voltage transfer function can be related to the scattering coefficients of the over-all network. The voltage transfer function at real frequencies is taken to be

$$\rho = \left|\frac{V_2}{V_1}\right| \quad (7)$$

where $|V_2|$ is the amplitude of the voltage appearing across the load, and $|V_1|$ is the open circuit (fixed) voltage amplitude of a normalized generator with unit internal impedance (pure resistance).

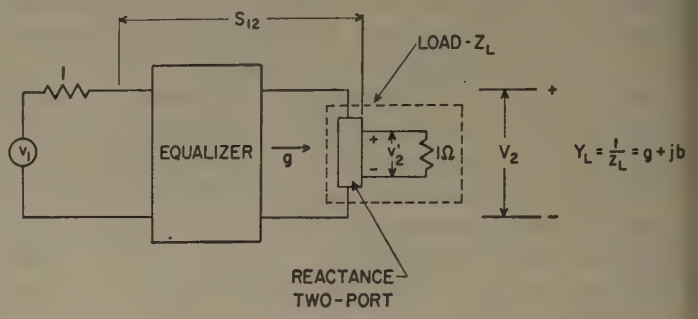


Fig. 1—Voltage transfer—general load.

The generator produces a voltage V_2' across the one ohm resistance in the Darlington representation of the load as a reactance two-port terminated in unit resistance as shown in Fig. 1. This voltage is related to V_1 by the voltage scattering function $S_{12}(p)$ of the over-all network (Fig. 1). Thus

$$V_2' = S_{12}(p) \frac{V_1}{2}. \quad (8)$$

Since the power at the input to the load is the same as that delivered to the one ohm resistor in the Darlington representation,

$$|V_2|^2 g(\omega) = |V_2'|^2 \quad (9)$$

where $g(\omega)$ is the input conductance of the load.

Combining (7), (8), and (9)

$$\rho^2 = \left|\frac{V_2}{V_1}\right|^2 = \frac{|S_{12}(j\omega)|^2}{4g(\omega)}. \quad (10)$$

Since $|S_{12}(j\omega)|^2 \leqq 1$, it is immediately clear that in any physical network

$$\rho^2 \leqq \frac{1}{4g(\omega)}. \quad (11)$$

In (10), $g(\omega)$ is specified by the load alone and ρ is directly proportional to $|S_{12}(j\omega)|$. Thus a necessary requirement for maximum voltage transfer is to maximize $|S_{12}(j\omega)|$ consistent with the general theorem on realizability given earlier. This is accomplished precisely as in the power transfer problem when $|S_{12}(j\omega)|$ and $|S_{11}(j\omega)|$ (prescribed) are related to $|S_{22}(j\omega)|$ by (3), (4), and (5). The general equation for optimum voltage transfer may then be written as

$$\rho^2 = \frac{(1 - |S_{22}(j\omega)|)(1 + |S_{11}(j\omega)|)}{4g(\omega)} \qquad (12)$$

$$|S_{22}(j\omega)| \geqq |S_{11}(j\omega)|.$$

The lossless and matched cases are then given as

$$\rho^2 = \frac{1 - |S_{22}(j\omega)|^2}{4g(\omega)} \text{ (lossless)} \qquad (13)$$

$$\rho^2 = \frac{1 - |S_{22}(j\omega)|}{4g(\omega)} \text{ (matched).} \qquad (14)$$

The integral equations for the latter two cases using (1) are

$$\frac{1}{2}\int_0^\infty f_i(\omega) \ln\left(\frac{1}{1 - 4\rho^2(\omega)g(\omega)}\right)\cdot d\omega = K_i \text{ (lossless)} \qquad (15)$$

$$\int_0^\infty f_i(\omega) \ln\left(\frac{1}{1 - 4\rho^2(\omega)g(\omega)}\right)\cdot d\omega = K_i \text{ (matched).} \qquad (16)$$

The weighting functions $f_i(\omega)$ and the parameters K_i are those tabulated in reference 3.

The only difference in form for the integral constraints in the two special cases is the factor $\frac{1}{2}$. However, since ρ appears under the integral sign, the solution for maximum scale factor is generally formed from a transcendental equation so that there is no direct relation between the voltage gain of lossless and matched equalizers even in the flat transfer case. In this latter case

$$\rho^2(\omega) = \begin{cases} C^2 & \omega_1 \leqq \omega \leqq \omega_2 \\ 0 & 0 \leqq \omega < \omega_1,\ \omega_2\ \omega \leqq \infty \end{cases} \qquad (17)$$

where C is the voltage gain constant to be maximized. The integrals (15) and (16) are then

$$\frac{1}{2}\int_{\omega_1}^{\omega_2} f_i(\omega) \ln\left(\frac{1}{1 - 4C^2 g(\omega)}\right)\cdot d\omega = K_i \text{ (lossless)} \qquad (18)$$

$$\int_{\omega_1}^{\omega_2} f_i(\omega) \ln\left(\frac{1}{1 - 4C^2 g(\omega)}\right)\cdot dw = K_i \text{ (matched).} \qquad (19)$$

In effect the problem of a flat voltage equalizer reduces to the solution of a power transfer problem where a non-flat power gain curve shape is specified.

B. Example—Flat Voltage Equalizer for R-L Load

As an example of a voltage equalizer problem, consider the case of a load consisting of the series combination of coil L and resistor R. The voltage transfer characteristic is to be a high pass one specified by:

$$\rho(x) = \begin{cases} C & x_c \leqq x \leqq \infty \\ 0 & 0 \leqq x < x_c \end{cases} \qquad (20)$$

where x is a normalized frequency variable and x_c is its cut off value:

$$x = \omega \frac{L}{R} \qquad (21)$$

$$x_c = \omega_c \frac{L}{R}. \qquad (22)$$

The load conductance is

$$g(x) = \frac{1}{R}\,\frac{1}{1 + x^2}. \qquad (23)$$

Since the load has only a simple zero of transmission at infinity, the weighting function $f(\omega)$ is unity and the integration constant K_i is given by:[3]

$$K \leqq \frac{\pi R}{L}. \qquad (24)$$

The possibility of attaining the equal sign in (24) is dictated by (11). If (23) is substituted in that equation then an upper bound is set on $\rho^2(x)$ for any value of x:

$$\rho^2(x) \leqq \frac{R(1 + x^2)}{4}. \qquad (25)$$

The permissible value of voltage gain increases with x, and thus for flat response

$$C^2 \leqq \frac{R(1 + x_c^2)}{4}, \qquad (26)$$

since the lowest permissible gain occurs at cut-off. The upper limit for $K = \pi R/L$ can only be attained if the value of C required in the integral equations does not violate (26). The integral relations given by (18) and (19) become:

$$A\frac{R}{L}\int_{x_c}^\infty \ln\frac{x^2 + 1}{x^2 + a^2}\,dx \leqq \frac{\pi R}{L} \qquad (27)$$

where

$$A = \begin{cases} 1 \text{ matched case} \\ \frac{1}{2} \text{ lossless case} \end{cases} \qquad (28)$$

and

$$a^2 = 1 - \frac{4C^2}{R}. \qquad (29)$$

For $a^2 \geqq 0$ integrating (27) gives

$$\pi(1 - a) - x_c \ln\frac{1 + x_c^2}{a^2 + x_c^2} - 2\tan^{-1} x_c + 2a\tan^{-1}\frac{x_c}{a} = \pi/A, \qquad a^2 \geqq 0. \qquad (30)$$

The equal sign is used in order to determine whether the value of a (hence C by (29)) exceeds the limit of (26). When $A = 1$, the only real solution for a in (30) occurs when $x_c = 0$. In that case:

$$a = a^2 = 0, \qquad C = \frac{R}{2}. \qquad (31)$$

The limit of (26) for $x_c = 0$ is also $C = R/2$, so that for flat transfer over $0 \leqq x \leqq \infty$, the maximum value per-

mitted by the integral constraint for a matched equalizer *can* be obtained and this flat gain is precisely the dc gain. A *lossless equalizer would give no gain advantage*, since the solution of (30) with $A=\frac{1}{2}$ results in a value of C exceeding that permitted by (26).

For values of $x_c>0$, negative values of a^2 are required to satisfy (30). Under these conditions the transcendental equation becomes:

$$\pi - x_c \cdot \ln \frac{x_c^2+1}{x_c^2-b^2} + b \ln \frac{x_c-b}{x_c+b} + 2\tan^{-1} x_c = \frac{\pi}{A}, \quad (32)$$

where

$$b^2 = -a^2 = \frac{4C^2}{R} - 1. \quad (33)$$

For values of $0 \leqq x_c \leqq 1.9$, the solution of (32) for C always exceeds that permitted by (26) for both values of A (lossless and matched cases). In this region, where the value of $\omega L/R$ is small at cut-off, the optimum flat gain is given by (26)

$$C = \sqrt{\frac{R(1+x_c^2)}{2}}, \qquad 0 \leqq x_c \leqq 1.9 \quad (34)$$

and the lossless equalizer gives no advantage in gain over the matched equalizer. For the medium range of $1.9 < x_c \leqq 4.8$, the flat gain of a matched equalizer as obtained from (32) is less than that given by (26), while the gain of the lossless equalizer is still limited by (26). Finally for the high range $4.8 < x_c \leqq \infty$, both lossless and matched equalizers have gains limited by the solution of (32).

When x_c is very large (32) is well approximated by:

$$b^2 + 1 = 2 + \frac{\pi x_c}{A}, \qquad x_c \gg 1. \quad (35)$$

Using (33), the optimum gain in this case is

$$\left.\begin{aligned} C &= \frac{\sqrt{R}}{2}\sqrt{2+\frac{\pi x_c}{A}} \\ &\cong \frac{1}{2}\frac{\pi L}{A}\omega_c \end{aligned}\right\} x_c = \frac{\omega_c L}{R} \gg 1. \quad (36)$$

Thus as the load resistance becomes negligible compared to the load reactance at cut-off, the ratio of maximum flat gain of lossless ($A=\frac{1}{2}$) and matched ($A=1$) equalizers becomes:

$$\frac{C\ (\text{lossless})}{C\ (\text{matched})} = \sqrt{2}. \quad (37)$$

The comparison of performance of lossless and matched equalizers for a flat, high-pass gain characteristic is summarized in graphical form in Fig. 2. The heavy line is the bounding curve defined by (26), and the dashed curves show the maximum flat gain of the lossless and matched equalizers as prescribed by the integral constraints.

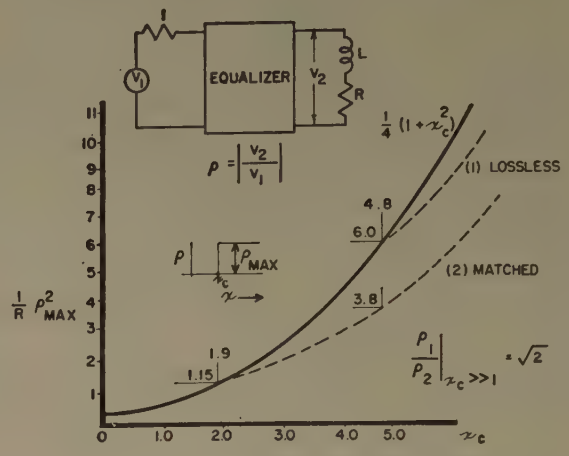


Fig. 2—Voltage equalizer *R-L* load.

Voltage Transfer Equalization of Reactive Load

The case of optimum voltage transfer from a finite generator to a reactive load is an important practical problem. However, the basic realizability theorem (theorem 1 quoted above) is not readily applicable. Accordingly, this equalization problem will be treated in a somewhat different way although the general point of view previously outlined will still be used. The optimum voltage equalization of an arbitrary lossless termination has not been considered in any complete fashion elsewhere, so that some details of the derivation of the realizability criteria (general constraints plus load constraints) will be given here. This will also serve to illuminate further the basis for the realizability theorem 1, since the two derivations parallel each other. It will be seen that, whereas the scattering coefficients were a natural tool for handling the equalization of a dissipative load, the open circuit impedance elements are more directly applicable to the reactive load problem.

In Fig. 3(a) on the following page, a finite generator of voltage V_1 and unit internal resistance is shown driving an equalizer terminated in an arbitrary reactance. The amplitude ratio of output load voltage to open circuit generator voltage is

$$\rho = \left|\frac{V_2}{V_1}\right| = |\bar{Z}_{12}(\omega)| \ \text{(general)} \quad (38a)$$

where $|\bar{Z}_{12}(\omega)|$ is the amplitude of open circuit transfer impedance of an over-all network shown in Fig. 3(b), consisting of the equalizer shunted at the generator side by a 1 ohm resistor and at the output side by the reactive load.

If the equalizer is designed to produce an input match, then the voltage ratio may be written as:

$$\rho = \left|\frac{V_2}{V_1}\right| = \frac{1}{2}\left|Z_{12}(\omega)\right| \text{ (matched equalizer)} \qquad (38)$$

where $|Z_{12}(\omega)|$ is the amplitude of the open circuit transfer impedance of the matching equalizer plus lossless load. (Generator resistance is not included.)

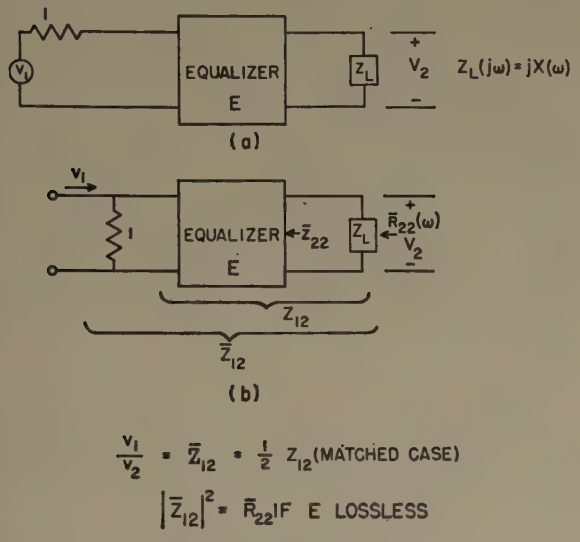


Fig. 3—Voltage transfer—reactive load.

If the equalizer is lossless, then in Fig. 3(b), with excitation at the load side of the over-all network, the power delivered here is equal to that dissipated in the 1-ohm shunting resistor. Thus

$$\rho^2 = |\bar{Z}_{21}(\omega)|^2 = |\bar{Z}_{12}(\omega)|^2 = \bar{R}_{22}(\omega) \text{ (lossless equalizer)} \qquad (39)$$

where $\bar{R}_{22}(\omega)$ is the real part of the open circuit output driving point impedance of the over-all network on $j\omega$:

$$\bar{R}_{22}(\omega) = \text{Re}\,\bar{Z}_{22}(j\omega). \qquad (40)$$

Since the equalization functions to be optimized [(38) and (39)] involve the open circuit impedance parameters, the requirements on over-all network physical realizability will be expressed in terms of these parameters. The necessary and sufficient conditions that an open circuit impedance matrix correspond to a physical two-port is that the matrix be positive real.[15] That is,

(a) $R_{11}(\omega)R_{22}(\omega) - R_{12}^2(\omega) \geqq 0, \quad R_{11}(\omega) \geqq 0, \qquad (41)$

(b) the open circuit impedances have no poles in the right half-plane,

(c) impedance element poles on the boundary be simple and the residues of $Z_{12}(p)$, $Z_{11}(p)$, and $Z_{22}(p)$ at these poles satisfy

$$a_{11}a_{22} - a_{12}^2 = 0, \qquad a_{11} \geqq 0.$$

Eq. (41) is the major physical constraint in the equalizer problem since if this is satisfied the remaining two requirements can in general be met by suitable design of the actual equalizer network.

The contraints of the load can be easily established by observing that the reactive load impedance $Z_L(p)$ is in parallel with the impedance seen looking in at the back of the equalizer. Thus

$$Z_{22} = \frac{z_{22}Z_L}{z_{22} + Z_L} \qquad (42)$$

$$\bar{Z}_{22} = \frac{\bar{z}_{22}Z_L}{\bar{z}_{22} + Z_L} \qquad (43)$$

where Z_{22} is the back-end open-circuit driving point impedance for the equalizer plus load with the 1-ohm resistor at the input removed, and $\bar{Z}_{22}$ is a similar quantity for the over-all network of Fig. 3(b). (z_{22} is back end equalizer impedance.)

Any zero of the reactive load must be simple and occur on $j\omega$. Further if Z_L is expanded in a power series at this zero, the first nonvanishing Taylor coefficient is positive.[16] Inspection of (42) and (43) show that the back end impedance Z_{22} or $\bar{Z}_{22}$ must vanish at this point and in the vicinity of the zero, $p_1 = j\omega_1$:

$$Z_{22}(p) = \bar{Z}_{22}(p) = Z_L(p) = a_1(p - p_1)\ p \to p_1. \qquad (44)$$

Eq. (44) is in fact entirely independent of the equalizer and its input termination.

The requirement that the back and impedance of the over-all network be constrained by (44) can be expressed in integral form by using the Cauchy formulas for Taylor coefficients. These take the form in the present instance of:

$$\frac{1}{2\pi j}\oint g_i(p)Z_{22}(p)dp = a_i \qquad (45)$$

where the closed path of integration is along $j\omega$ (avoiding by small semi-circular indentations any boundary poles) and is completed around the semicircle of infinite radius enclosing the right-half p plane. $g_i(p)$ are the weighting functions to select the appropriate Taylor coefficient of the load, a_i. If the contributions of the small indentations are accounted for, and $g_i(p)$ is modified slightly to be even in ω along $j\omega$, (45) may be written in terms of $R_{22}(\omega)$ as follows:

[15] C. M. Gewertz, "Synthesis of a finite, four-terminal network from its prescribed driving-point functions and transfer function," *Jour. Math. & Physics*, vol. 12, pp. 1–257; 1932–33. Also "Network Synthesis," Williams and Williams Book Co., Baltimore, 1933.

[16] This follows from Foster's reactance theorem. R. M. Foster, "A reactance theorem," *Bell. Sys. Tech. Jour.*, vol. III, pp. 259–267; April, 1924.

$$\int_0^\infty \frac{1}{\omega^2} R_{22}(\omega)d\omega = \frac{\pi a_0}{2} - \pi \sum_m \frac{b_m}{\omega_m{}^2} \text{ (zero at } p = 0) \quad (46)$$

$$\int_0^\infty R_{22}(\omega)d\omega = \frac{\pi a_\infty}{2} - \pi \sum_m b_m \text{ (zero at } p = \infty) \quad (47)$$

$$\int_0^\infty \frac{1}{(\omega^2 - \omega_1{}^2)^2} R_{22}(\omega)d\omega = \frac{\pi a_1}{4\omega_1{}^2} - \pi \sum_m \frac{b_m}{(\omega_m{}^2 - \omega_1{}^2)^2} \text{ (zero at } p = \pm j\omega_1) \; \omega_m \neq \omega_1. \quad (48)$$

In every case b_m is the residue of any Z_{22} poles on $j\omega$ and is always positive. Such poles can only reduce the permissible integration constant. The only other possibility of reducing the right hand sides of (46), (47), (48) is for z_{22} (or $\bar{z}_{22}$) of the equalizer to have a zero coincident with Z_L. Since these relations must all be satisfied simultaneously by Z_{22} (or $\overline{Z}_{22}$), it may be necessary to introduce elements in the equalizer which cause coincident zeros and/or additional poles on $j\omega$. Another point which should be mentioned here is that, in referring to (44), Eqs. (45), (46), (47), (48) are *valid both for the barred and unbarred quantities.*

Finally, since any zero of the load is a zero of voltage transfer, the following physical realizability theorem may be stated:

Theorem 2:

The necessary and sufficient conditions for a combined network consisting of a two-port and a prescribed reactive termination to be physically realizable are

(a) The combined network must have a positive real open circuit impedance matrix.

(b) All zeros of the reactive load impedance must be contained in the open circuit transfer impedance of combined network. (Zeros are all simple.)

(c) The integral constraints given by (46), (47), (48) and summarized by the form

$$\int_0^\infty f_i(\omega)R_{22}(\omega)d\omega = K_i \quad (49)$$

must be simultaneously satisfied by the back end resistance of the combined network at each zero of the load.

The similarity between this theorem and theorem 1 is obvious.

Flat Gain Equalization of a Reactive Load

A. Form of Integral Constraints

As an example of the application of theorem 2, the case of flat voltage transfer for a prototype low-pass equalizer defined by:

$$\rho = \left|\frac{V_2}{V_1}\right| = |\overline{Z}_{12}(\omega)| = \begin{cases} \overline{C} & \text{for } 0 \leqq \omega \leqq 1 \\ 0 & \text{for } 1 < \omega \leqq \infty \end{cases} \quad (50)$$

will be examined.

In the case of the matched equalizer, condition (a) of theorem 2 leads to

$$R_{22}(\omega) - R_{12}{}^2(\omega) \geqq 0 \text{ (matched equalizer).} \quad (51)$$

This is obtained from (41) with $R_{11}(\omega) \equiv 1$, $0 \leqq \omega \leqq \infty$. Now

$$R_{12}(\omega) = |Z_{12}(\omega)| \cos(\phi(\omega) + \theta(\omega)) \quad (52)$$

where ϕ is the minimum phase characteristic of the combined network (equalizer and load), and $\theta(\omega)$ any phase characteristic obtained with a tandem combination of unit voltage transfer networks, considered as part of the equalizer (i.e. the latter need not be a minimum phase network).

The function $\phi(\omega)$ is directly found graphically or analytically from the shape of the $Z_{12}(\omega)$ amplitude function and is not affected by the scale factor on this function.[1] $\theta(\omega)$ is independent of $\phi(\omega)$ and is continuous with positive slope.[1] If (52) is substituted into (51), it is clear that a necessary condition for maximum $|Z_{12}(\omega)|$ is that the equal sign be used. Thus

$$R_{22}(\omega) = |Z_{12}(\omega)|^2 \cos^2(\phi + \theta) = R_{12}{}^2(\omega). \quad (53)$$

This means that the combined matched equalizer-load network will require no more than one resistor.[15] Suppose that $|Z_{12}(\omega)|$ is represented as

$$|Z_{12}(\omega)| = \begin{cases} C & 0 \leqq \omega \leqq 1 \\ \dfrac{C}{\omega^k} & 1 < \omega \leqq \infty \end{cases} \quad (k \text{ a positive integer}) \quad (54)$$

where

$$C = 2\overline{C}. \quad (55)$$

As $k \to \infty$, the transfer characteristic defined by (54) approaches the gain shape given by (50) (it differs by the factor $\frac{1}{2}$).

The minimum phase characteristic defined by (54) is given below: (See reference 1, pp. 312–318)

$$\phi(\omega) = \frac{k}{\pi}\int_1^\infty \log \coth \frac{|u|}{2}\, du = \begin{cases} k\phi_1(\omega) & 0 \leqq \omega \leqq 1 \\ k\phi_2(\omega) & 1 < \omega \leqq \infty. \end{cases} \quad (56)$$

In this equation $u = \log \Omega/\omega$, where Ω is the running frequency variable of integration. The definite integral is of course a function of ω.

The approximate values of $\phi_1(\omega)$ and $\phi_2(\omega)$ are given below for reference only. They are not used in the derivation, and merely indicate the nature of these functions. They are continuous and monotonic.

$$\phi_1(\omega) \leqq -k\frac{2\omega}{\pi} \qquad 0 \leqq \omega \leqq 1 \quad (56a)$$

$$\phi_2(\omega) \leqq -k\frac{\pi}{2} \qquad 1 < \omega \leqq \infty. \quad (56b)$$

The integral (49) of theorem 2 for the *matched* flat voltage equalizer becomes

$$C^2\int_0^1 f_i(\omega)\cos^2[k\phi_1(\omega)+\theta(\omega)]d\omega + C^2\int_1^\infty \frac{f_i(\omega)}{\omega^k}\cos^2(k\phi_2(\omega)+\theta(\omega))d\omega = K_i. \quad (57)$$

The second integral clearly approaches zero as k becomes infinite. Expanding the squared cosine term, the first integral may be written as:

$$\frac{1}{2}\int_0^1 f_1(\omega)d\omega+\frac{1}{2}\int_0^1[f_i(\omega\cos 2\theta(\omega)]\cos 2k\phi_1(\omega)d\omega + \frac{1}{2}\int_0^1[f_i(\omega)\sin 2\theta(\omega)]\sin 2k\phi_1(\omega)d\omega. \quad (58)$$

Now let $\gamma(\omega)=2\phi_1(\omega)$; hence $\gamma'(\omega)\,d\omega=d\gamma$, $\omega=h(\gamma)$, so that the second and third integrals above become:

$$\frac{1}{2}\int_{\gamma(0)=0}^{\gamma(1)} g_1(\gamma)\cos k\gamma d\gamma + \frac{1}{2}\int_{\gamma(0)=0}^{\gamma(1)} g_2(\gamma)\sin k\gamma d\gamma$$

where

$$g_1(\gamma)=\frac{f_i(\gamma)\cos 2\theta(\gamma)}{\gamma'(\gamma)}, \quad g_2(\gamma)=\frac{f_i(\gamma)\sin 2\theta(\gamma)}{\gamma'(\gamma)}. \quad (59)$$

The first of these integrals is essentially the kth order Fourier coefficient for a function defined equal to $g_1(\gamma)$ in the interval $\gamma(0)$ to $\gamma(1)$, periodic for all γ and even in γ. The second is a kth order Fourier coefficient for a function defined as equal to $g_2(\gamma)$ over the interval $\gamma(0)$ to $\gamma(1)$, periodic and odd in γ. As k becomes large, the Fourier coefficients approach zero, hence these integrals vanish in the limit $k\to\infty$.

The final representation of the integral constraint for a matched equalizer with the low-pass flat voltage gain of (50) is therefore

$$2\overline{C}^2\int_0^1 f_i(\omega) = K_i \text{ (matched flat low-pass case).} \quad (60)$$

A similar derivation for a flat band-pass over the interval ω_1 to ω_2 (zero gain elsewhere) gives

$$2\overline{C}^2\int_{\omega_1}^{\omega_2} f_i(\omega) = K_i \text{ (matched flat band-pass case).} \quad (60a)$$

Both (60) and (60a) are entirely independent of whether minimum or nonminimum phase networks are used, but apply only to a flat voltage gain response of amplitude $\overline{C}$ over a finite interval with zero gain elsewhere.

Returning to the low-pass case, let

$$\int_0^1 f_i(\omega)d\omega = A_i. \quad (61)$$

Eq. (60) may then be written

$$\overline{C}^2=\frac{K_i}{2A_i} \text{ (matched).} \quad (62)$$

This equation is independent of whether minimum or nonminimum phase networks are used.

For a lossless equalizer, (39) may be used directly in connection with (50) and placed into the expression of theorem 2. Thus the lossless equalizer with flat transfer characteristics given by (50) must satisfy:

$$\int_0^\infty f_i(\omega)\overline{R}_{22}(\omega)d\omega = \int_0^\infty f_i(\omega)\,|\,\overline{Z}_{12}(\omega)\,|^2 d\omega = \overline{C}^2\int_0^1 f_i(\omega)d\omega \quad (63)$$

or

$$\overline{C}^2=\frac{K_i}{A_i} \text{ (lossless case)} \quad (64)$$

and the ratio of optimum matched and lossless equalizer gains for *the flat transfer case* and reactive load is:

$$\frac{\overline{C}\text{ (lossless)}}{\overline{C}\text{ (matched)}}=\sqrt{2}. \quad (65)$$

B. *Example of a C-LC Load*

As an example of the application of the material given above, suppose it is required to design a flat voltage equalizer over the band $0\leqq\omega\leqq 1$ for a three-element load. The load consists of a capacitor C' in shunt with a series circuit of coil L and condenser C.

This load impedance has zeros at $p=\infty$, and $p=\pm j\omega_1$ with $\omega_1=1/LC$.[17]

The Taylor coefficients at these zeros are:

$$(p=\infty)a_\infty=\frac{1}{C'} \quad (66)$$

$$(p=\pm j\omega_1)a_1=\frac{2L}{C}. \quad (67)$$

Referring to (47) and (48) and neglecting the pole residue terms,

$$K_1=\frac{\pi a_\infty}{2}=\frac{\pi}{2C'} \quad (68)$$

$$K_2=\frac{\pi a_1}{4{\omega_1}^2}=\frac{\pi}{2}L^2. \quad (69)$$

For the zero at infinity the weighting function is unity and

$$A_1=\int_0^1 d\omega = 1. \quad (70)$$

For the zero at $\pm j\omega_1$, the weighting function is

$$f_2(\omega)=\frac{1}{(\omega^2-{\omega_1}^2)^2}$$

and

[17] Presumed outside of pass-band.

$$A_2 = \int_0^1 \frac{d\omega}{(\omega^2 - \omega_1^2)^2} = \frac{1}{2\omega_1^2(\omega_1^2 - 1)} + \frac{1}{4\omega_1^3} \cdot \ln \frac{\omega_1 + 1}{\omega_1 - 1} \cdot \tag{71}$$

Consider a specific numerical case with

$$C' = \frac{1}{2}$$
$$L = C = \frac{1}{\sqrt{2}} \cdot \tag{72}$$

For these conditions the various constants may be evaluated to give:

$$\frac{K_1}{A_1} = \pi \tag{73}$$

$$\frac{K_2}{A_2} = \frac{\pi/4}{0.35} = \frac{\pi}{1.40} \cdot \tag{74}$$

The only way to satisfy these constraints simultaneously is to reduce the value of K_1 by using parasitic elements in the equalizer. This can be done for example by use of a shunt condenser across the equalizer output and in parallel with the load of value $C'' = 0.2$. As a result *the maximum gain of this equalizer is not limited by the capacitance* C', but by the series LC circuit, and the maximum value of this gain for a low-pass matched equalizer is

$$\overline{C} = \sqrt{\frac{\pi}{2.80}} \text{ (matched)}, \tag{75}$$

and for a lossless equalizer the maximum gain is

$$\overline{C} = \sqrt{\frac{\pi}{1.40}} \text{ (lossless)}. \tag{76}$$

It is up to the designer to decide whether a matched input to the equalizer system is worth sacrificing for a 40 per cent increase in gain.

C. Design of a Finite Equalizer Network

In the previous discussion ideal characteristics were assumed for the equalizer amplitude response. The voltage gain was presumed absolutely flat in the pass-band, and the cut-off was taken as infinitely sharp. In order to apply theorem 2 to the design of a finite network, a gain characteristic may be assumed in analytic form (i.e. an even rational function of ω which is always positive). The analysis in such a case involves determination of the R_{22} characteristic and its substitution in the integral equations to determine the maximum gain scale factor. To illustrate this procedure a simple example will be given for the voltage equalization and match of a unit capacitive termination.[18] A Butterworth type of low-pass response is assumed for a matched equalizer

$$|Z_{12}(\omega)|^2 = \frac{C^2}{1 + \omega^4} \cdot \tag{77}$$

The amplitude of $Z_{12}(\omega)^2$ is down by $\frac{1}{2}$ at $\omega = 1$, which is considered the normalized cut-off frequency. C is a gain constant maximum value of which is to be determined by applying theorem 2.

Eq. (77) may be factored to give the open circuit transfer impedance $Z_{12}(p)$ of the required combined matched equalizer-load network. This function must have no right half-plane poles.

$$Z_{12}(p) = \frac{C}{p^2 + \sqrt{2}\,p + 1} \cdot \tag{78}$$

The impedance $Z_{12}(p)$ has the zero of the load at infinity as prescribed by theorem 2 because of the particular choice of amplitude function in (77).

The real component of $Z_{12}(p)$ along $p = j\omega$ is

$$R_{12}(\omega) = C\frac{1 - \omega^2}{1 + \omega^2} \tag{79}$$

and, applying (53) for the optimum matched equalizer,

$$R_{22}(\omega) = R_{12}^2(\omega) = C^2 \frac{(1 - \omega^2)^2}{(1 + \omega^4)^2} \cdot \tag{80}$$

The complex impedance $Z_{22}(p)$ may be obtained by expanding $R_{22}(\omega)$ in partial fractions as described on pages 204–205 of reference 1. The result[19] is

$$Z_{22}(p) = C^2 \frac{0.353p^3 + p^2 + 1.065p + 1}{p^4 + 2.82p^3 + 4p^2 + 2.82p + 1} \cdot \tag{81}$$

This function has the load zero at infinity as prescribed by theorem 2.

The integral constraint [Eq. (47)] gives for unit capacitive load:

$$\int_0^\infty R_{22}d\omega = C^2 \int_0^\infty \frac{(1 - \omega^2)^2}{(1 + \omega^4)^2} \cdot d\omega = \frac{\pi}{2} \cdot \tag{82}$$

If the definite integral is evaluated, the gain factor C is found to be

$$C = 1.68 \tag{83}$$

or

$$\overline{C} = \frac{C}{2} = 0.84 \text{ (matched case)}.[20] \tag{84}$$

Maximum gain for completely ideal low-pass characteristic (in matched case) obtained from (62) is

$$\overline{C} \text{ (ideal)} = 0.89 \text{ (matched case)} \tag{85}$$

[18] This example is taken from a thesis by L. Norde for the degree M.E.E. at the Polytechnic Institute of Brooklyn, June, 1953.[7]

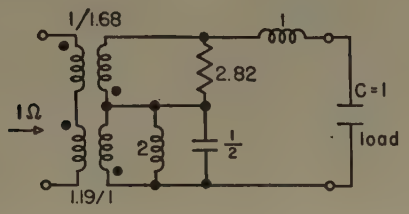
[19] In general, if the resulting $Z_{22}(p)$ function *does not* contain the required load zeros, it is necessary to form new rational functions for $|Z_{12}(\omega)|^2$ (possibly by the use of common numerator and denominator factors) until *this requirement is satisfied.*

[20] This constraint could have been directly obtained from the network synthesis by choosing $\overline{C}$ to give the prescribed load capacitance

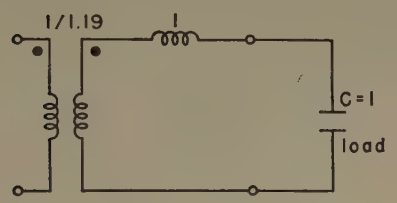
so that the simple characteristic of (77) is a reasonable compromise.

The final matched equalizer plus load is specified by (78), (81) and

$$Z_{11}(p) = 1.$$



(a) MATCHED EQUALIZER A=0.84



(b) LOSSLESS EQUALIZER A=1.19

Fig. 4—Equalizers for capacitive load and voltage gain $=A/\sqrt{1+w^4}$.

The resultant network is shown in Fig. 4(a). Observe that the load capacitance is removable leaving a physical equalizer, and only one resistor appears in the network.

The design of a *lossless* equalizer from a specified $|Z_{12}(\omega)|^2$ characteristic which satisfies theorem 2 is straightforward. The equalizer network is merely synthesized from the back end as a reactance two-port terminated in a 1-ohm resistor by the Darlington procedure for the special case of an infinite impedance generator.[10] The 1 ohm resistor (input generator impedance) and load are then removed, and the remaining lossless network is the required equalizer.

For the example of (77) with

$$|\overline{Z}_{12}(\omega)|^2 = \frac{\overline{C}^2}{1+\omega^4} = R_{22}(\omega), \tag{86}$$

the integral requirement gives

$$\int_0^\infty \frac{\overline{C}^2}{1+\omega^4}\,d\omega = \frac{\pi}{2}$$

and

$$\overline{C} = 2^{0.25} = 1.19 \text{ (lossless equalizer)}. \tag{87}$$

The network is shown in Fig. 4(b).

Directional Electromagnetic Couplers*

BERNARD M. OLIVER†, FELLOW, IRE

Summary—**The natural coupling between parallel transmission lines is inherently directional. Very simple and cheap directional couplers can be made which utilize this effect. By introducing appropriate variation of coupling with distance a wide variety of transmission characteristics may be realized, including high-pass (ideally, infinite bandwidth) characteristics. The coupled lines may be of equal or different impedance. Thus, a simple configuration of four wires (or two wires and ground) can serve simultaneously as a directional coupler, filter, and transformer.**

Introduction

THE PROPAGATION of waves on systems of parallel conductors was investigated many years ago[1] in connection with the problem of crosstalk between open wire lines and between cable pairs. These early investigations were directed toward eliminating the natural coupling rather than using it. Later investigations[2–4] using matrix algebra provided elegant mathematical treatments of the general multiconductor transmission problem. The generality of these treatments obscures the simplicity of the special cases we are interested in here. When directional couplers became popular a host of papers[5–7] appeared on the subject. With few exceptions[5] the emphasis in these papers was on waveguide devices. The purpose of this paper is to provide a reasonably simple and not unnecessarily general treatment of the directional coupling action of ordinary transmission lines.

The following assumptions will be made:

1. The loss is negligible; i.e. $\alpha l \ll 1$ for any l considered.
2. The conductor sizes and spacings are infinitesimal compared with the wavelength.
3. All currents are confined to the conductor surfaces.
4. The medium is homogeneous and isotropic (except where controlled anisotropy is introduced to give co-directional coupling).
5. The lines are balanced and the coupling is symmetrical.

Assumptions (1) to (4) are readily met in the 1 to 1,000 mc range; and assure that only TEM waves need be considered. All such waves propagate at the velocity of light in the medium.

Assumption (5) means that only balanced modes will exist; and limits the configurations to those shown in Fig. 1 (opposite page). The coupling between two unbalanced (single-wire-to-ground) circuits is a degenerate case of Fig. 1(a) so all results save those involving cross-coupling apply to this case as well.

Normal Modes and Line Constants

The propagation of waves on two parallel lines such as shown in Fig. 1 is described by the set of differential equations:

$$\left.\begin{aligned} \frac{\partial e_1}{\partial z} + L_{11}\frac{\partial i_1}{\partial t} + L_m\frac{\partial i_2}{\partial t} &= 0 \\ \frac{\partial i_1}{\partial z} + C_{11}\frac{\partial e_1}{\partial t} - C_m\frac{\partial e_2}{\partial t} &= 0 \\ \frac{\partial e_2}{\partial z} + L_{22}\frac{\partial i_2}{\partial t} + L_m\frac{\partial i_1}{\partial t} &= 0 \\ \frac{\partial i_2}{\partial z} + C_{22}\frac{\partial e_2}{\partial t} - C_m\frac{\partial e_1}{\partial t} &= 0 \end{aligned}\right\} \tag{1}$$

where:

L_{jj}, C_{jj} $(j=1, 2)$ are the self inductance and capacitance per unit length of line j, in the presence of line k $(k=2, 1)$ taking e_k and i_k to be zero,
L_m is the mutual inductance per unit length,
C_m is the mutual capacity per unit length.

As in any system involving coupled linear circuits, we have a set of simultaneous linear differential equations which admits a particularly simple set of solutions: the *normal modes* of the system. In this case we have four normal modes (two in each direction) characterized by fixed ratios of e_2/e_1 and i_2/i_1 which do not change as the wave propagates.[8] The modes are characterized by:

$$\frac{e_2}{e_1} = \pm\sqrt{\frac{L_{22}}{L_{11}}} = \pm\sqrt{\frac{C_{11}}{C_{22}}} \tag{2}$$

$$e_1 i_1 = e_2 i_2. \tag{3}$$

The choice of the plus sign corresponds to "lateral" excitation in Fig. 1(a), the minus sign to "diagonal" excitation. In either normal mode equal power is associated with each line.

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[1] J. R. Carson and R. S. Hoyt, "Propagation of periodic waves over a system of parallel wires," *Bell Sys. Tech. Jour.*, vol. IV, no. 3, pp. 495–545; July, 1927.

[2] L. A. Pipes, "Matrix theory of multiconductor transmission lines," *Phil. Mag.*, vol. 24, pp. 97–100; 1937.

[3] S. Koizumi, "Mehrpolleitungstheorie," *Arch. für Electrotech.*, vol. 33, pp. 171–188, 609–622; 1939.

[4] S. O. Rice, "Steady state solution of transmission line equations," *Bell Sys. Tech. Jour.*, vol. XX, no. 2, pp. 131–178; April, 1941.

[5] A. A. Pistolkors and M. S. Neumann, "An instrument for direct measurement of the travelling wave coefficient in feeders," *Electrosvyaz*, vol. IX, no. 4; April, 1941. (RTP trans. no. 1525.)

[6] F. Bolinder, "Approximate theory of the directional coupler," PROC. I.R.E., vol. 39, p. 291; March, 1951.

[7] For a listing of about ninety directional coupler references see R. F. Schwartz, "Bibliography on directional couplers," *Trans. I.R.E.*, vol. MTT-2, no. 2, pp. 58-63; July, 1954.

[8] That is to say there is no "crosstalk" between normal modes. As Carson and Hoyt, *loc. cit.*, point out, if each telephone channel were assigned a normal mode, transposition of telephone wires would be unnecessary. Transposition merely confines each normal mode to a particular pair.

Upon making these substitutions and letting

$$k_C = \frac{C_M}{\sqrt{C_{11}C_{22}}}, \qquad k_L = \frac{L_M}{\sqrt{L_{11}L_{22}}} \tag{4}$$

(1) reduces to the familiar transmission line equations:

$$\frac{\partial e_i}{\partial z} + L_{ii}(1 \pm k_L)\frac{\partial i_i}{\partial t} = 0$$
$$\frac{\partial i_i}{\partial z} + C_{ii}(1 \mp k_C)\frac{\partial e_i}{\partial t} = 0 \qquad i = 1, 2.$$

The velocity of propagation

$$v = \frac{1}{\sqrt{L_{ii}C_{ii}(1 \pm k_L)(1 \mp k_C)}}$$

must be the same for all modes, i.e. whether the upper signs or the lower signs are chosen. This can only be true if $k_L = k_C = k$. Hence for natural isotropic coupling

$$\frac{L_M}{C_M} = \sqrt{\frac{L_{11}L_{22}}{C_{11}C_{22}}} \tag{5}$$

and

$$v = \frac{1}{\sqrt{L_{ii}C_{ii}(1 - k^2)}}.$$

Further, since v must be independent of k, $L_{ii}C_{ii}$ must vary as $(1-k^2)^{-1}$. Both L and C will change with k, but the essential variation will be due to C alone, the remaining variations being mutually reciprocal and negligible except when the conductors are in close proximity

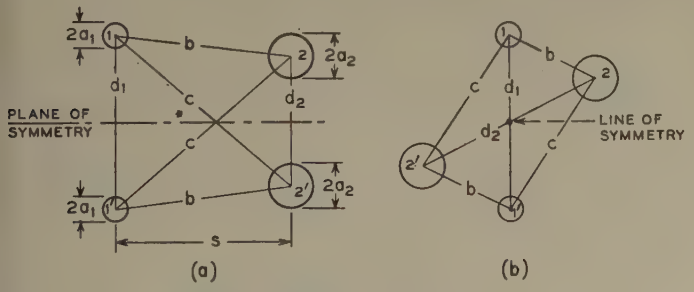

Fig. 1—Coupled balanced lines.

With reasonable conductor separations:

$$k = \frac{\log \frac{c}{b}}{\sqrt{\log \frac{d_1}{a_1} \log \frac{d_2}{a_2}}} \tag{6}$$

$$C_{ii} = \frac{\pi\epsilon}{(1 - k^2)\log \frac{d_i}{a_i}} = \frac{C_i}{(1 - k^2)} \tag{7}$$

$$L_{ii} = \frac{\mu}{\pi} \log \frac{d_i}{a_i} = L_i. \tag{8}$$

C_i and L_i $(i = 1, 2)$ are the self-capacity and inductance of either line in the absence of the other.

When the lines are identical (i.e. if $a_1 = a_2$, $d_1 = d_2$) (6) may be written:

$$k = \frac{\log \sqrt{1 + \left(\frac{d}{s}\right)^2}}{\log \frac{d}{a}}. \tag{6a}$$

For small coupling $[(d/s) \ll 1]$ tl is simplifies to

$$k \cong \frac{d^2}{2s^2} \frac{1}{\log \frac{d}{a}} \tag{6b}$$

which, for lines in air becomes

$$k \cong \frac{60}{Z_0}\left(\frac{d}{s}\right)^2.$$

Natural Directional Coupling

Assume a wave is traveling from left to right in line 1 and consider the induced voltages and currents produced in an elemental length dz of line 2 terminated in impedances Z_a and Z_b as shown in Fig. 2. The rising voltage

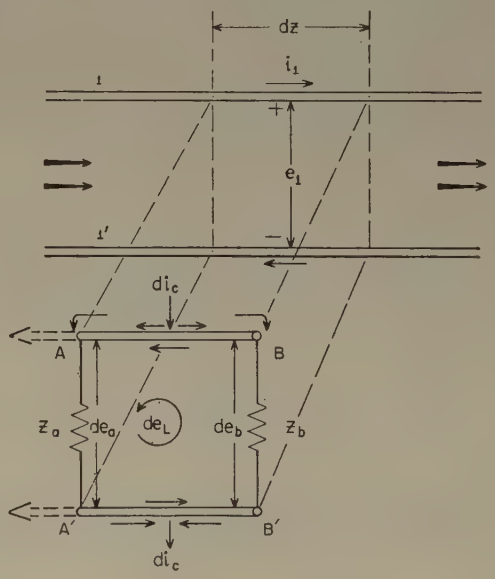

Fig. 2—A coupled elemental length of line.

on line 1 will produce the displacement current $di_C = (\partial/\partial t)(C_M e_1)dz$ while the increasing current in line 1 will produce the emf $de_L = (\partial/\partial t)(L_M i_1)dz$ around the loop as shown. The total voltage across terminals BB' will be

$$de = \frac{Z_a Z_b}{Z_a + Z_b} di_C - \frac{Z_b}{Z_a + Z_b} de_L$$
$$= \frac{Z_b}{Z_a + Z_b} \frac{\partial}{\partial t}(Z_a C_M e_1 - L_M i_1)dz.$$

This voltage will be zero if

$$Z_a = \frac{L_M}{C_M} \frac{i_1}{e_1}.$$

But

$$\frac{i_1}{e_1} = \sqrt{\frac{C_{11}}{L_{11}}},$$

while from (5)

$$\frac{L_M}{C_M} = \sqrt{\frac{L_{11}L_{22}}{C_{11}C_{22}}},$$

so that we require

$$Z_a = \sqrt{\frac{L_{22}}{C_{22}}}.$$

Since this is the characteristic impedance of line 1 in the presence of line 2, the voltage and current at BB' will still be zero if the conductors AB and $A'B'$ are extended an arbitrary distance to the left and then terminated in Z_a. This follows since the induction in each added elemental section produces no voltages or currents at its right-hand terminals. The entire induced voltage and current in each elemental section appear at the left and are of the proper relative polarity to launch a wave traveling to the left. Any voltage or current appearing at the extreme right is a result of reflection. We may conclude, therefore, that the natural coupling between parallel lines is inherently directional. In particular it is *contra-directional:* the induced wave travels in the opposite direction to the inducing wave.

Obviously if either the electric or magnetic coupling were eliminated the coupling would be nondirectional, while if either were reversed the coupling would be *co-directional.* This latter case will be considered later.

If $Z_a = Z_{22}$, we find, since all the voltage and current appear at $A-A'$:

$$\left.\begin{aligned} de_A = de_L &= \frac{\partial}{\partial t}(L_M i_1)dz = ik\sqrt{LC}\,\frac{\partial e_1}{\partial t}dz \\ di_A = di_C &= \frac{\partial}{\partial t}(C_M e_1)dz = ik\sqrt{LC}\,\frac{\partial i_1}{\partial t}dz \end{aligned}\right\}. \quad (9)$$

Contra-Directional Coupler with Constant Coupling

From here on we will assume that the two coupled lines are of equal impedance. Little will be lost by this since all results may be converted to the unequal case merely by multiplying by the appropriate impedance ratio. Let:

L_0, C_0, Z_0 = Inductance, capacitance, impedance of either line in the absence of the other ($k=0$).

L, C, Z = Corresponding quantities in the presence of other line.

L_m, C_m = Mutual inductance and capacitance per unit length.

Z^+ = Ratio of voltage to current in one line for "plus" or "lateral" mode.

Z^- = Ratio of voltage to current in one line for "minus" or "diagonal" mode.

k = Coupling coefficient.

We then have the following relations

$$\left.\begin{aligned} &L \cong L_0, \quad C \cong \frac{C_0}{1-k^2}, \quad Z = Z_0\sqrt{1-k^2} \\ &Z^+ = Z\sqrt{\frac{1+k}{1-k}} = Z_0(1+k) \\ &Z^- = Z\sqrt{\frac{1-k}{1+k}} = Z_0(1-k) \end{aligned}\right\}. \quad (10)$$

Now consider two lines with constant coupling over a length l, as shown in Fig. 3. If the connecting lines are of impedance $Z = \sqrt{L/C}$ (*not* $Z_0 = \sqrt{L_0/C_0}$) and are properly terminated, then from the previous results, we should expect a wave incident on terminals $a-a'$ to produce an output at $b-b'$ and at $c-c'$, but none at $d-d'$. In analyzing the behavior of this coupler it is convenient to represent the input E_0 at $a-a'$ as the sum of an input $E_0/2$ applied to $a-a'$ and $c-c'$ in the $+$mode, and an input $E_0/2$ applied to these same terminals in the $-$ mode. The two normal mode inputs cancel at $c-c'$ but add at $a-a'$ to give the actual input.

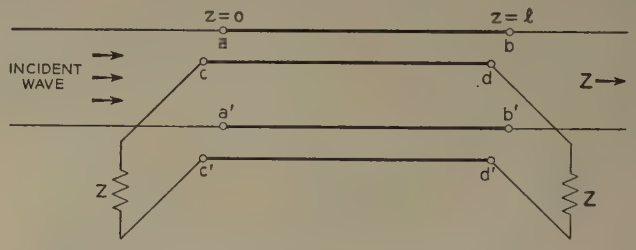

Fig. 3—A directional coupler.

If both normal mode inputs are supplied over uncoupled lines of impedance Z then at the input terminals a reflection will be produced. If

$$\zeta = \sqrt{\frac{1+k}{1-k}}$$

then for the $+$ mode the reflection coefficient will be

$$\rho^+ = \frac{\zeta - 1}{\zeta + 1} = \rho$$

while for the $-$ mode

$$\rho^- = \frac{\frac{1}{\zeta} - 1}{\frac{1}{\zeta} + 1} = \frac{1-\zeta}{1+\zeta} = -\rho.$$

The two reflections cancel at the terminals $a-a'$, add to produce an output at $c-c'$. A complete picture of the

TABLE I

Time and Place	Wave	+ Mode	− Mode	Total Line 1	Total Line 2
$t=0$ $z=0$	Inc	$\frac{1}{2}$	$\frac{1}{2}$	1	0
	Refl	$\frac{1}{2}\rho$	$-\frac{1}{2}\rho$	0	ρ
	Trans	$\frac{\zeta}{\zeta+1}$	$\frac{1}{\zeta+1}$	1	ρ
$t=\tau$ $z=l$	Inc	$\frac{\zeta}{\zeta+1}$	$\frac{1}{\zeta+1}$	1	ρ
	Refl	$-\rho\frac{\zeta}{\zeta+1}$	$\rho\frac{1}{\zeta+1}$	$-\rho^2$	$-\rho$
	Trans	$\frac{2\zeta}{(\zeta+1)^2}$	$\frac{2\zeta}{(\zeta+1)^2}$	$\frac{4\zeta}{(\zeta+1)^2}$	0
$t=2\tau$ $z=0$	Inc	$-\rho\frac{\zeta}{\zeta+1}$	$\rho\frac{1}{\zeta+1}$	$-\rho^2$	$-\rho$
	Refl	$\rho^2\frac{\zeta}{\zeta+1}$	$\rho^2\frac{1}{\zeta+1}$	ρ^2	ρ^3
	Trans	$-\rho\frac{2\zeta}{(\zeta+1)^2}$	$\rho\frac{2\zeta}{(\zeta+1)^2}$	0	$-\rho\frac{4\zeta}{(\zeta+1)^2}$
$t=3\tau$ $z=l$	Inc	ρ^2 times row 4. Hence geometric series of outputs produced from here on.			

Fig. 4—Sequence of reflections and transmissions with unit impulse input. Quantities in heavy boxes are first direct output and first two coupled outputs.

Note: $\zeta=\sqrt{\frac{1+k}{1-k}}$, $\rho=\frac{\zeta-1}{\zeta+1}$.

action may be had by taking the component normal mode inputs to be impulses of area $\frac{1}{2}$. An impulse thus appears at $c-c'$ the moment the actual input enters the coupled region. The transmitted impulses propagate down the coupler to produce a combined output at $b-b'$ but none (total) at $d-d'$. The reflections from this end propagate back to $z=0$ to produce an output at $c-c'$ of reversed polarity, together with further reflections to the right. At this point the conditions duplicate those just after the impulse first entered the coupler, except that all amplitudes are reduced by the factor

$$\left(\frac{\zeta-1}{\zeta+1}\right)^2.$$

As a result there now follows a geometric series of impulse outputs at $b-b'$ and $c-c'$, each impulse smaller than the last by the factor

$$\left(\frac{\zeta-1}{\zeta+1}\right)^2$$

and delayed an additional time $2\tau=2l/v$.

The initial series of reflected and transmitted impulse amplitudes for both normal modes and for their sums are given in Table I (Fig. 4) left, while the actual outputs are shown in Fig. 5 for various values of k.

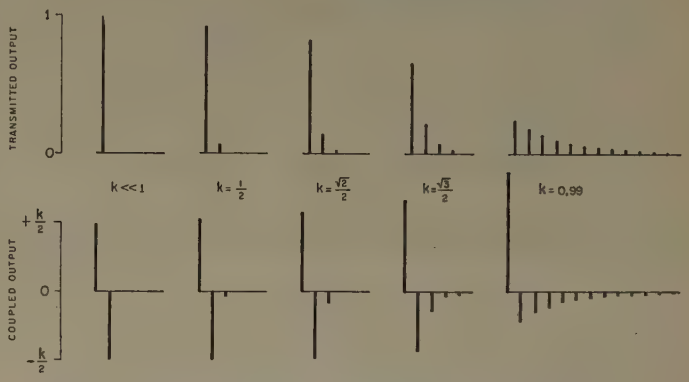

Fig. 5—Impulse response of contra-directional coupler with constant coupling.

Since the spectrum of a unit impulse occurring at a time τ is $e^{-i\omega t}$, the spectrum of the output for a unit impulse input (*i.e. the frequency response*) may readily be found from the preceding. Thus, for the direct, or transmitted, output (at $b-b'$) we have

$$G=\frac{4\zeta}{(\zeta+1)^2}e^{-i\omega\tau}\left[1+\rho^2e^{-i2\omega\tau}+\rho^4e^{-i4\omega\tau}+\cdots\right]$$
$$=\frac{4\zeta}{(\zeta+1)^2}\,\frac{e^{-i\omega\tau}}{1-\rho^2e^{-i2\omega\tau}}$$
$$=\frac{\sqrt{1-k^2}}{\sqrt{1-k^2}\cos\beta l+i\sin\beta l}. \qquad (11)$$

Likewise the series of impulses at $c-c'$ gives for the coupled output

$$F=\rho-\rho\frac{4\zeta}{(\zeta+1)^2}e^{-i2\omega\tau}\left[1+\rho^2e^{-i2\omega\tau}+\rho^4e^{-i4\omega\tau}+\cdots\right]$$
$$=ik\frac{\sin\beta l}{\sqrt{1-k^2}\cos\beta l+i\sin\beta l}. \qquad (12)$$

From (11) and (12) we note that

$$\frac{F}{G}=i\frac{k}{\sqrt{1-k^2}}\sin\beta l,$$

so that for any k the two outputs are in quadrature and their ratio varies sinusoidally with ω or l. This quadrature property is characteristic of all directional couplers. We further note that $|G|^2+|F|^2=1$ as must be true since the coupler is lossless.

Fig. 6 (page 1690) shows the magnitude and phase of F as a function of βl for various values of k. It will be seen that for small k the transmission varies sinusoidally with frequency (or length) and the phase linearly. With

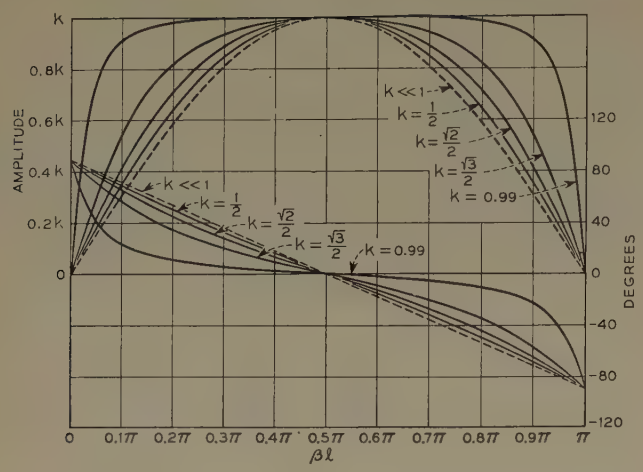

Fig. 6—Contra-directional coupler with constant coupling, k. With tighter coupling the transmission is flatter and the midband delay is less.

$k \ll 1$ the coupler covers a 3 to 1 band between 3 db points. With larger values of k the transmission is even broader; for example $k = \sqrt{2}/2$ (which produces an equal power split at $\beta l = \pi/2$) gives a characteristic covering a 4 to 1 band. *The maximum transmission* to the coupled line always occurs when $l = (2n-1)(\lambda/4)$, and *has the magnitude k*. With a harmonic input of amplitude E_0 applied to terminals $a-a'$, the voltages as a function of distance along the coupler are

$$e_1 = E_0 \frac{\sqrt{1-k^2}\cos\beta(l-z) + i\sin\beta(l-z)}{\sqrt{1-k^2}\cos\beta l + i\sin\beta l}$$

for the voltage on line 1, and

$$e_2 = E_0 ik \frac{\sin\beta(l-z)}{\sqrt{1-k^2}\cos\beta l + i\sin\beta l}$$

for the voltage in line 2 (the branch line). Fig. 7 shows the magnitudes of e_1 and e_2 for $k = 1/\sqrt{10}$.

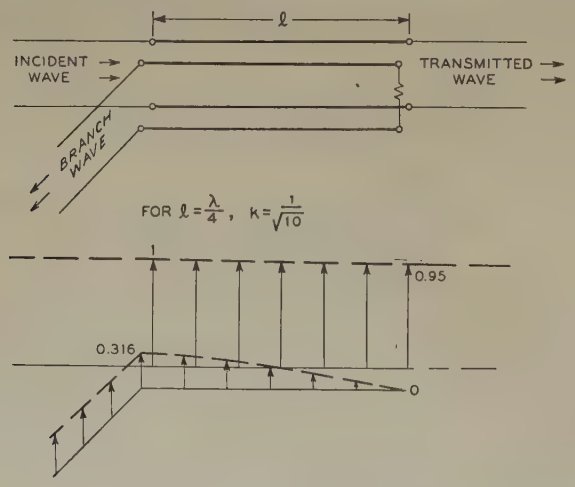

Fig. 7—Illustrating the voltage magnitudes in a contra-directional coupler.

Variable Coupling—Filter Properties

If the coupling coefficient, k, is varied with distance along the coupler, a wide variety of transmission characteristics may be realized. In order to avoid reflections which would destroy the directivity, the impedance, Z, of the lines must be kept constant as $k(z)$ changes. This requires that Z_0 vary as $[1-k^2(z)]^{-1/2}$. However, if the maximum coupling is such that $k^2(z) \ll 1$ and if extremely high directivities are not required, this complication is unnecessary.

If $|F(\omega)|^2 \ll 1$ for all frequencies, the reaction of the induced waves on the inducing waves may be neglected and the transmission of the coupler is readily calculated. By integrating (9) after including the delay from the input terminals to z and back to the output terminals, we find:

$$F = i\beta \int k(z) e^{-i2\beta z} dz.$$

Upon substituting vt/z for z and ω/v for β, this becomes

$$F(\omega) = i\omega \int \frac{1}{2} k\left(\frac{vt}{2}\right) e^{-i\omega t} dt. \tag{13}$$

If both sides of (13) are divided by $i\omega$, the left-hand side will be recognized as the spectrum of the response to a step function, and the right-hand side as the Fourier transform (spectrum) of $\frac{1}{2}k(vt/2)$. Thus, *the step function response is one half the coupling function $k(z)$ with $vt/2$ substituted for z*, i.e.

$$s(t) = \frac{1}{2} k\left(\frac{vt}{2}\right). \tag{14}$$

Further, since the impulse response, $f(t)$, is the time derivative of the step response we have

$$f(t) = \frac{v}{4} k'\left(\frac{vt}{2}\right). \tag{15}$$

Suppose now we wish to realize a coupler having a specified transmission $F(\omega)$ between input and coupled output. The corresponding impulse and step responses are

$$f(t) = \frac{1}{2\pi} \int F(\omega) e^{i\omega t} d\omega.$$

$$s(t) = \frac{1}{2\pi} \int \frac{F(\omega)}{i\omega} e^{i\omega t} d\omega.$$

From (14) and (15) we then find

$$k'(z) = \frac{4}{v} f\left(\frac{2z}{v}\right). \tag{16a}$$

$$k(z) = 2s\left(\frac{2z}{v}\right). \tag{16b}$$

If $F(0) \neq 0$ then $s(\infty) \neq 0$ and from (16b) a coupler of infinite length is required. Aside from this, any $F(\omega)$ which can be realized as an ordinary filter characteristic can be realized in a directional coupler. If $s(t)$ reverses sign the same must be true of $k(z)$. This requires the conductors of either line to be transposed.

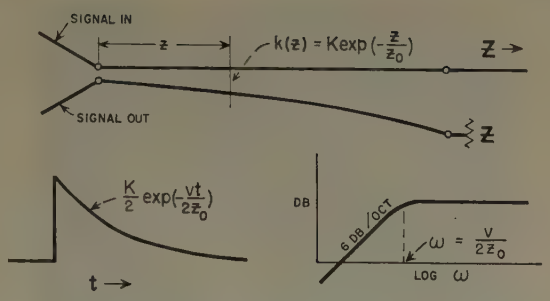

Fig. 8—Exponential variation of coupling with distance. Reversing coupler end-for-end reverses transient response and phase characteristic; leaves gain unchanged.

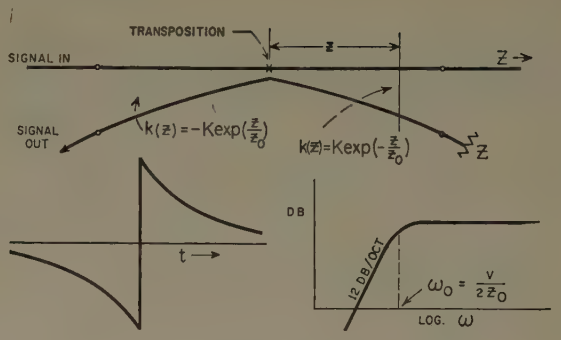

Fig. 9—Symmetrical high-pass coupler.

Thus filter characteristics whose step-function responses reverse sign can be realized only in balanced-line structures. Sharp cut-off filters cannot be realized in two dimensional coaxial structures where such transposition is impossible. Figs. 8 and 9 show two high-pass couplers together with their corresponding step-function responses and steady-state transmissions. Actually, high-frequency cutoffs will be present because the coupling cannot in practice be initiated or reversed infinitely sharply. However, with small lines closely spaced, 100 to 1 bandwidths should be feasible.

Besides changes in Z (which can be avoided) as k changes there is another source of backward transmission (i.e. output at terminals $d-d'$) in a contra-directional coupler. As the conductor separation is changing so that the lines are at an angle θ with respect to each other, the electric coupling in the length dz is the same as for $\theta=0$, but the magnetic coupling is decreased by the factor $\cos\theta$. The two no longer cancel in the backward direction, and a small backward transmission proportional to $\frac{1}{2}(1-\cos\theta)\approx\theta^2/4$ will result. This is most apt to happen at the ends, where the connecting lines diverge; thus, whenever possible, the coupling should gradually be reduced to low values at the ends.

Co-Directional Couplers

By adding excess capacitance either as capacitors along the diagonals c in Fig. 1(a), or as dielectric slabs along the sides c in Fig. 1(b) the capacitive coupling may be reversed, to give $k_C=-k_L$. The required capacitance per unit length, ΔC, for each diagonal (or side) is

$$\Delta C=\frac{4k}{1-k}C=\frac{4k}{1-k}\frac{C_0}{1-k^2}.$$

The resulting effective capacitance per unit length for each line is

$$C'=C+\frac{\Delta C}{2}$$

$$C'=C\frac{1+k}{1-k}=\frac{C_0}{(1-k)^2}.$$

With this loading the impedance for both normal modes is the same

$$Z=Z^+=Z^-=\sqrt{\frac{L}{C'}}=(1-k)\sqrt{\frac{L_0}{C_0}},$$

but the velocities of propagation are different. The phase constants are

$$\beta^+=\beta(1+k),$$
$$\beta^-=\beta(1-k),$$

where

$$\beta=\omega\sqrt{LC'}=\frac{\omega\sqrt{L_0C_0}}{1-k}.$$

Thus the $-$ mode (which is unloaded) propagates at the velocity of light while the $+$ mode is slowed by the factor $(1-k)/(1+k)$. An input applied to $a-a'$ can be decomposed as before into two equal normal-mode components which add at $a-a'$, cancel at $c-c'$. As these propagate down the lines the modes will slip in relative phase till the cancellation takes place in line 1 and the addition in line 2. The power simply transfers back and forth from line to line. With an input $E_0e^{i\omega t}$ to line 1 we find

$$\frac{e_1}{E_0}=e^{-i\beta z}\cos k\beta z$$

$$\frac{e_2}{E_0}=-ie^{-i\beta z}\sin k\beta l.$$

With a coupler of length l, and constant coupling, the direct transmission (terminals $b-b'$) is therefore

$$G=e^{-i\beta l}\cos k\beta l \qquad (17)$$

while the coupled transmission (terminals $d-d'$) is

$$F=-ie^{-i\beta l}\sin k\beta l. \qquad (18)$$

In co-directional couplers, variation of k merely alters the rate at which energy transfer takes place with distance. The output still varies sinusoidally with frequency, k in (17) and (18) being replaced by the average coupling

$$\bar{k}=\frac{1}{l}\int_0^l k(z)dz.$$

No interesting filter properties can be produced, for the

increments from each elemental length merely add without relative delay, and the Fourier relations do not apply. The chief point of interest about co-directional couplers is that *complete* power transfer takes place when $k\beta l = (2n-1)\pi/2$, no matter how small k itself may be.

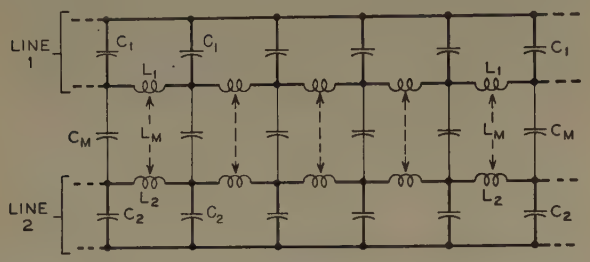

Fig. 10—Coupled lumped lines. With such lines directional couplers can be made to operate at video frequencies.

Coupled Lumped Lines and Helixes

By coupling the coils of two lumped constant delay lines, and by adding capacitance between the nodes as shown in Fig. 10, directional coupling may be achieved at video frequencies. All that is required is that

$$L_M = \pm k\sqrt{L_1 L_2}$$

where

$$k = \frac{C_M}{\sqrt{(C_1 + C_M)(C_2 + C_M)}}.$$

In lumped lines the magnetic coupling can be reversed by reversing the winding sense of the coils on either line, so that here either co-directional or contra-directional coupling may be realized. The same situation exists in the case of coupled helixes. By winding one helix over the other in the opposite sense, co-directional coupling with complete power transfer may be achieved. The velocities of propagation differ because the magnetic fields add for the − mode to give greater inductance than for the + mode for which the fields cancel. (Or one can say that while the natural coupling is contra-directional in the fine, i.e. wire-to-wire, the opposite winding sense makes both induced and inducing waves travel in the same axial direction.)

An Experimental VHF Coupler

Fig. 11 shows a double contra-directional coupler designed for reflectometer measurements in the vhf band. The conductors are No. 12 wire, spaced ½ inch on centers, and the two lines are separated ½ inch in the center sections, 1 inch at the ends. Each coupler is 2 feet long. The computed forward loss at four points is given below, in Table II. These figures agree fairly closely with the measured results shown in Fig. 12. The directivity in this case was limited by the terminations which were ordinary ½ w 300Ω resistors.

TABLE II

Freq.	TV Channel	Loss
54 mc	2	23 db
88 mc	5	20 db
177 mc	7	17.5 db
216 mc	13	19 db

Fig. 11—Experimental vhf directional couplers.

If high directivities are not imperative, and if forward losses on the order of 20 db are permissible, merely running two lengths of twin lead side-by-side for a quarter-wavelength will provide a satisfactory coupler. The directivity is impaired by the dielectric which, by increasing C but not C_m, makes $k_C < k_L$.

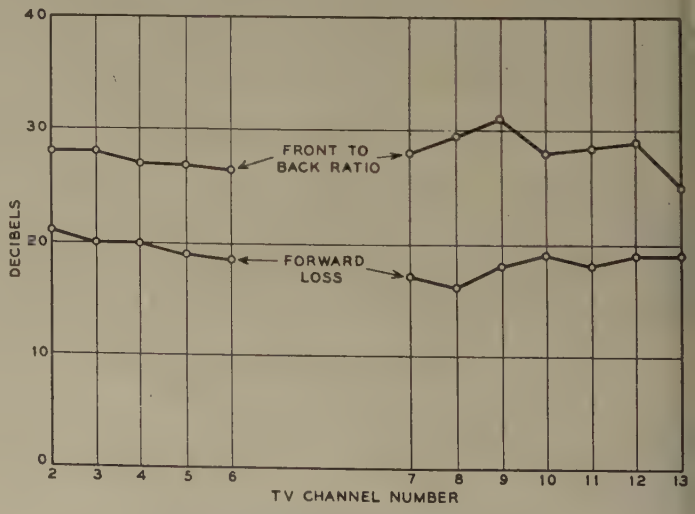

Fig. 12—The performance of a simple balanced 300Ω directional coupler in the vhf TV bands.

Co-directional couplers can also be made with twin lead using the configuration of Fig. 1(b), with the dielectric along sides c. When adjusted to make $k_C = -k_L$, sides b will be slightly shorter than sides c; i.e. the two twin leads will be separated slightly less than their width. The input and output connections at both ends are made across the two diagonals.

Generation of Standard Fields in Shielded Enclosures*

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Summary—**In this paper a method for calibrating loop antennas in a shielded enclosure is analyzed. The scheme considered involves the generation of a known field in the vicinity of the loop by an energized wire mounted near the ceiling of the enclosure. The published formulas for computing the field are based upon simplifying approximations which result in large inaccuracies in many cases. To extend the usefulness of the method new formulas are developed with which the field can be predicted with better than 1 per cent accuracy in the low and medium frequency ranges for most installations. The results can be applied over most of the high frequency region with a possible reduction of accuracy. A method is presented for estimating the error in this region. Finally, the passible limitation of the method at very low frequencies is discussed.**

Introduction

THE "transmission line" method for generating a known magnetic field inside a shielded enclosure is employed extensively for calibrating loop antennas.[1–3] The system as it is commonly used is illustrated in Fig. 1. The loop to be calibrated and the radiating line shown are mounted in a plane equidistant from the side walls; the resistive termination is chosen

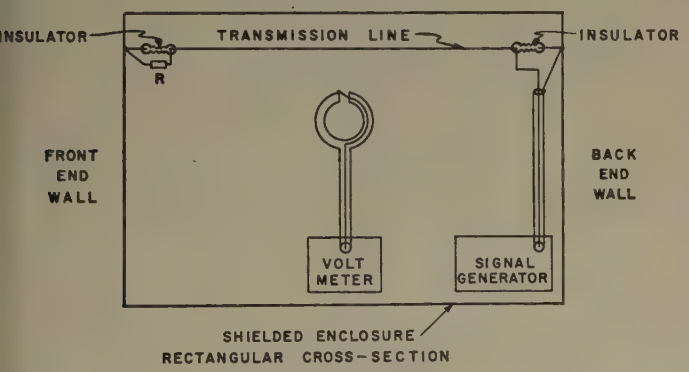

Fig. 1—Conventional transmission line setup.

to minimize standing waves on the line. With such an arrangement a predictable magnetic field is generated in the vicinity of the loop for determining the loop sensitivity. Formulas for computing the field intensity under the line appear in the literature. Bond[1] considers the system to be equivalent to an infinitely long wire mounted under a perfectly conducting plane of infinite extent (the plane of the ceiling). The formulas given by him are derived from the Biot-Savart law for stationary fields applied to the wire and to its image in the conducting plane. Terman and Pettit,[2] and Sandretto,[3] employ a similar solution, adding, however, an image in the plane of the floor.

These formulas are inexact even in the medium frequency range. In order for the boundary conditions to be satisfied at the surface of the conducting walls, an infinite set of images is required. Thus a question arises regarding the error involved in the calculation when only one or two images are considered. An additional question arises regarding the frequency limitation of these formulas. In the high frequency range the assumption of quasi-stationary conditions may result in excessive error while at very low frequencies the image method, which assumes perfect conductivity, may not be valid.

The analysis presented in this paper leads to a formula for the magnetic field under the wire to a predictable accuracy in the lower and medium frequency ranges. In most applications the accuracy is well within 1 per cent. The analysis is based upon the assumption that quasi-stationary conditions prevail. However, to determine the validity of such a result in the high frequency region an analysis is made of the field of the wire and its images taking retardation time into account. The assumption made here is that no standing waves appear on the wire. This development leads to a method for estimating the error incurred by using the quasi-stationary formula in the high frequency region. Finally, the limitation of the image method at very low frequencies is discussed and a method for estimating the error is obtained.

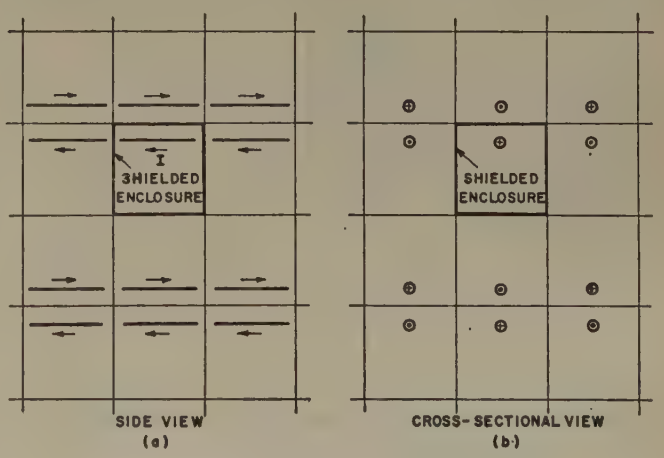

Fig. 2—Images due to current flowing in transmission line. (a) Side-view. (b) Cross-sectional view.

Field Under the Wire Assuming Quasi-Stationary Conditions

At those frequencies at which propagation time along the wire and through space can be ignored, the Biot-Savart law applies. Therefore:

$$H_\theta = \frac{I_0 \sin \omega t}{2\pi\rho} \tag{1}$$

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[1] D. S. Bond, "Radio Direction Finders," McGraw-Hill Book Company, Inc., New York, N. Y., pp. 221–229; 1944.

[2] F. E. Terman and J. M. Pettit, "Electronic Measurements," McGraw-Hill Book Company, Inc., New York, N. Y., pp. 405–406; 1952.

[3] P. C. Sandretto, "Principles of Aeronautical Radio Engineering," McGraw-Hill Book Company, Inc., New York, N. Y., pp. 140–142; 1942.

where H_θ is the magnetic field intensity in the direction of increasing θ in a cylindrical coordinate system, ρ, θ, z, with the wire along the z-axis.

The three dimensional set of images required for a complete solution of the problem is shown in Fig. 2 (preceding page). At these low frequencies the current at all points along the wire and its images in the end wall, shown in Fig. 2(a), will be in phase at any given instant of time. Thus, the field inside the enclosure reduces to one due to a two dimensional set of conductors, each of infinite length, carrying in-phase current. Fig. 3 shows the two dimensional array of images applicable in this this case; the field due to the wire and each image is obtained by use of (1). The conductors will be treated in pairs as is shown by the dotted rectangles in the figure. Consider a pair at position (m, n) carrying a current $I = \pm I_0 \sin \omega t$. The component of the field in a direction perpendicular to the side walls at a point P under the wire due to this pair is:

$$H_{mn} = (-1)^m \frac{I}{2\pi}\left[\frac{x_n - d}{(x_n-d)^2 + w_m^2} - \frac{x_n + d}{(x_n+d)^2 + w_m^2}\right] \quad (2)$$

where the symbols are defined in Fig. 3.

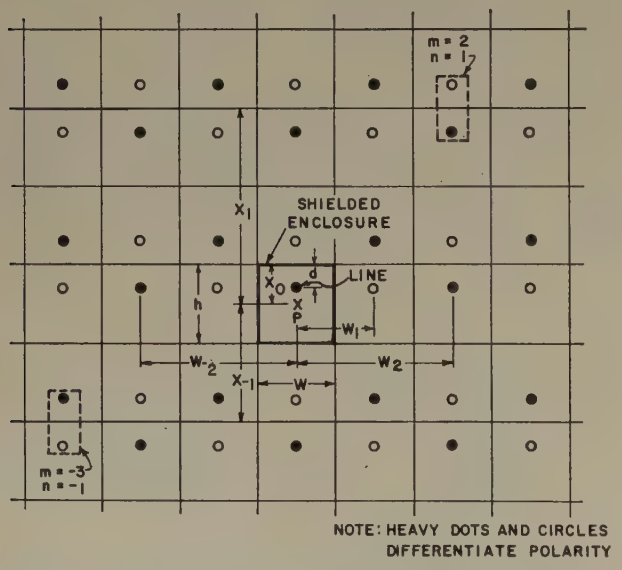

Fig. 3—Expanded view of images in cross-sectional plane.

From the figure it is also seen that

$$x_n = x_0 + 2nh$$

$$w_m = mw.$$

Using these relationships and summing the effects of all the conductors, as illustrated on Fig. 3,

$$H_{\text{total}} = \frac{I}{2\pi} \sum_{n=-\infty}^{\infty} \sum_{n=-\infty}^{\infty} (-1)^m \left[\frac{x_0 + 2nh - d}{(x_0+2nh-d)^2 + m^2w^2} - \frac{x_0 + 2nh + d}{(x_0+2nh+d)^2 + m^2w^2}\right]. \quad (3)$$

Eq. (3) can be reduced to a one dimensional series by application of the known series

$$\pi z \cot \pi z = 1 + 2z^2 \sum_{m=1}^{\infty} \frac{1}{z^2 + m^2}. \quad (4)$$

The result obtained is

$$H_{\text{total}} = \frac{I}{2\pi} \sum_{n=-\infty}^{\infty} \frac{\pi}{w}\left[\frac{2 \sinh \pi d/w}{\cosh (x_0+2nh)\pi/w - \cosh \pi d/w} - \frac{2 \sinh 2\pi d/w}{\cosh 2(x_0+2nh)\pi/w - \cosh 2\pi d/w}\right]. \quad (5)$$

It can be shown[4] that the following are upper and lower bounds, respectively, to the series (5).

$$H_{\text{total upper}} = \frac{I}{w}\left\{\left[\frac{\sinh \alpha}{\cosh \beta - \cosh \alpha} - \frac{\sinh 2\alpha}{\cosh 2\beta - \cosh 2\alpha}\right] + \frac{2 \tanh \alpha \cosh \alpha}{\exp (2\gamma) - 1}\left[\frac{1}{\exp \beta - 2 \exp (-2\gamma) \cosh \alpha} + \frac{1}{\exp (-\beta) - 2 \exp (-2\gamma) \cosh \alpha}\right] - \frac{2 \tanh 2\alpha \cosh 2\alpha}{\exp (4\gamma) - 1} \cdot \left[\frac{1}{\exp 2\beta} + \frac{1}{\exp (-2\beta)}\right]\right\} \quad (6)$$

$$H_{\text{total lower}} = \frac{I}{w}\left\{\left[\frac{\sinh \alpha}{\cosh \beta - \cosh \alpha} - \frac{\sinh 2\alpha}{\cosh 2\beta - \cosh 2\alpha}\right] + \frac{2 \tanh \alpha \cosh \alpha}{\exp (2\gamma) - 1}\left[\frac{1}{\exp \beta} + \frac{1}{\exp (-\beta)}\right] - \frac{2 \tanh 2\alpha \cosh 2\alpha}{\exp (4\gamma) - 1} \cdot \left[\frac{1}{\exp 2\beta - 2 \exp (-4\gamma) \cosh 2\alpha} + \frac{1}{\exp (-2\beta) - 2 \exp (-4\gamma) \cosh 2\alpha}\right]\right\} \quad (7)$$

where $\alpha = \pi d/w$
$\beta = \pi x_0/w$
$\gamma = \pi h/w.$

Inspection of those terms which differ in (6) and (7) will show that in most installation (with h/w of the order unity or greater) the difference between the bounds is considerably less than 1 per cent. In an unusual installation in which $h/w = 0.5$ the difference between the bounds may be as high as 4 per cent for a particularly poor choice of the other parameters. A convenient approximate solution which has a value between the two bounds is given by:

[4] Moore School of Elec. Eng., Res. Div. Rep. No. 54-04, pp. 78–83; July 15, 1953.

$$H_{\text{total}} = \frac{I}{w}\left\{\frac{\sinh\alpha}{\cosh\beta-\cosh\alpha} - \frac{\sinh 2\alpha}{\cosh 2\beta-\cosh 2\alpha} + 4\sinh\alpha\cosh\beta/[\exp(2\gamma)-1] - 4\sinh 2\alpha\cosh 2\beta/[\exp(4\gamma)-1]\right\}. \quad (8)$$

A further simplification can be made by ignoring the last term in (8). For most practical applications, this term will be found to contribute very little to the total result. Also, if we include the approximation $1/[\exp(2\gamma)-1] \cong \exp(-2\gamma)$ which is satisfactory for $h/w > 1/2$,

$$H_{\text{total}} = \frac{I}{w}\left[\frac{\sinh\alpha}{\cosh\beta-\cosh\alpha} - \frac{\sinh 2\alpha}{\cosh 2\beta-\cosh 2\alpha} + 4\sinh\alpha\cosh\beta\exp(-2\gamma)\right]. \quad (9)$$

Eqs. (8) and (9) give the field at a point, usually selected to be at the center of the loop. If the loop area is large compared to the distance of the loop from the wire, the average value of field over the loop area is required. This can be obtained by an integration of (8) or (9). In the integration to follow only the first two terms of (8) or (9) are treated, since for the other terms, which are due to images far removed from the loop, the average value and the value at the center are practically

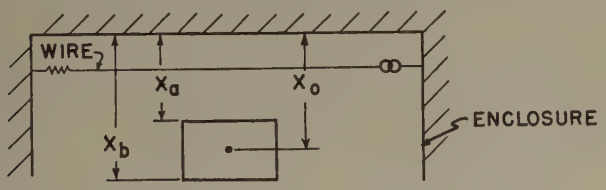

Fig. 4—Rectangular loop of large area under the line.

identical. Refer to Fig. 4 which shows a rectangular loop mounted under the wire. The field varies only in the vertical direction, hence:

$$\overline{H} = \frac{1}{(x_b - x_a)}\int_{x_a}^{x_b} H_{\text{total}}\,dx \quad (10)$$

where $\overline{H}$ is the average field over the loop area. Using (9) and (10) as discussed above, the average field intensity over the loop becomes

$$\overline{H} = \frac{I}{2\pi(x_b-x_a)}\ln_e\left\{\frac{\sinh^2[(x_b-d)\pi/2w]}{\sinh^2[(x_a-d)\pi/2w]}\cdot\frac{\sinh^2[(x_a+d)\pi/2w]}{\sinh^2[(x_b+d)\pi/2w]}\cdot\frac{\sinh[(x_a-d)\pi/w]}{\sinh[(x_b-d)\pi/w]}\cdot\frac{\sinh[(x_b+d)\pi/w]}{\sinh[(x_a+d)\pi/w]}\right\} + \frac{4I}{w}\cosh(\pi x_0/w)\sinh(\pi d/w)\exp(-2\pi h/w). \quad (11)$$

In many cases the loop encountered is circular, requiring the average value over a circular area. A method suggested by Bond is to obtain the average field over an inscribed square on the circle and over a circumscribed square on the circle. These are upper and lower bounds to the average field. These two results may, in turn, be averaged to obtain a reasonable approximation to the average field. It has been suggested that a square loop of area equal to that of the circular loop may be substituted for the circle for the purposes of calculation.[5]

A comparison of the result obtained by the use of (7) and the result obtained by the use of a "one image equation" is instructive. Assuming one volt applied to a line terminated in 395 ohms (2.532 ma), $h = 110$ inches, $w = 126$ inches, $d = 12$ inches, $x_0 = 51.4$ inches, the field calculated using (9) is 0.0687 volts per meter. Calculating the field using the same parameters but using a formula which considers only one image in the ceiling the field strength obtained is 0.0575 volts per meter. These values were obtained by multiplying the magnetic field by the impedance of free space, 120π ohms. The formula developed here thus gives a result which is about 20 per cent higher than the formula based on one image for this choice of parameters. If an image in the plane of the floor is added, as suggested in some references, the result would be 0.0957 volt per meter. The result established in this paper is about 28 per cent below this figure.

High Frequency Analysis

The effect of propagation time will now be determined. Essentially, this involves obtaining an expression similar to (1) which includes the effect of the finite velocity of both the current along the wire and of the field through space. The expression obtained is then used to estimate the error incurred when the results of the previous section are applied to predict a field in the high frequency range.[6]

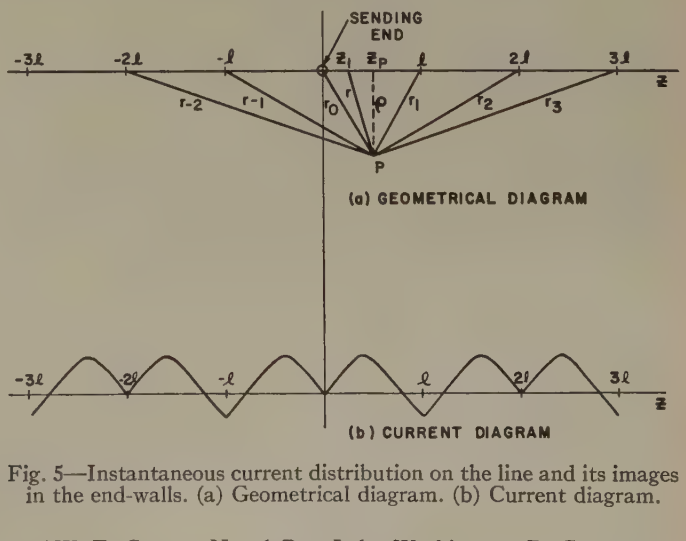

Fig. 5—Instantaneous current distribution on the line and its images in the end-walls. (a) Geometrical diagram. (b) Current diagram.

[5] W. E. Garner, Naval Res. Lab., Washington, D. C., personal communication.

[6] A more rigorous expression can be derived by summing contributions from the two dimensional grid of Fig. 2 to obtain an expression for the field intensity applicable over a wider range of frequencies. Such a summation is indeed possible using the results of this section, but its utility would be limited due to the complexity of the net result.

The following is an evaluation of the field due to the current in the line and its images in the end walls as indicated in Fig. 5. The wire itself is represented by the segment of the z-axis from the origin to l. The images in the end walls are represented by the segments l to $2l$, $2l$ to $3l$, etc., and $-l$ to $-2l$, $-2l$ to $-3l$, etc. If a current of the form $I_0 \sin \omega t$ is initiated at the origin and is assumed to travel to the right with no loss and no reflection, then at any point $0<z<l$ the current is $I_0 \sin \omega(t-z/c)$ where c is the velocity of propagation. The velocity of propagation in the copper conductors is equal to the free space velocity. With the aid of Fig. 5(b), the following can be written for the instantaneous current at any point, z_1 along the z-axis.

For $0 < z < l$, $I = I_0 \sin \omega(t - z_1/c)$

For $-l < z < 0$, $I = I_0 \sin \omega(t + z_1/c)$

For $l < z < 2l$, $I = I_0 \sin \omega[t - (2l - z_1)/c]$

For $-2l < z < -l$, $I = I_0 \sin \omega[t - (2l + z_1)/c]$

For $2l < z < 3l$, $I = I_0 \sin \omega[t + (2l - z_1)/c]$

For $-3l < z < -2l$, $I = I_0 \sin \omega[t + (2l + z_1)/c]$

For $3l < z < 4l$, $I = I_0 \sin \omega[t - (4l - z_1)/c]$

For $-4l < z < -3l$, $I = I_0 \sin \omega[t - (4l + z_1)/c]$

In general:

$$I = I_0 \sin \omega\left(t - \frac{al}{c} + \frac{bz_1}{c}\right) \tag{12}$$

where: a takes on even integer values 0, ± 2, ± 4, etc. and b takes on values ± 1, as indicated above.

The field at point P in Fig. 5(a) can be computed using the retarded vector potential

$$A^z = \frac{1}{4\pi} \int_L \frac{I(t - r/c)}{r}\, dz_1, \tag{13}$$

where L indicates integration along the appropriate conductor, and the relation

$$H_\theta = -\partial A^z/\partial \rho \tag{14}$$

where H_θ is in the direction of increasing θ in a cylindrical co-ordinate system ρ, θ, z, with the wire along the z-axis. Using (12), (13) and (14)

$$H_\theta = -\frac{1}{4\pi} \frac{\partial}{\partial \rho} \int_L \frac{I_0 \sin \omega[t - (r + al - bz_1)/c]}{r}\, dz_1. \tag{15}$$

The solution to this integral can be shown to be[7]

$$H_\theta = \frac{I_0 \rho}{4\pi} \cdot \frac{\sin \omega|t - (r + al - bz_1)/c]}{rb[r - b(z_1 - z_p)]}\Bigg|_L \tag{16}$$

where z_p is as indicated in Fig. 5. If the appropriate limits are applied and H_θ summed over all the conductors indicated in Fig. 5 the following result is obtained:

[7] Moore School of Elec. Eng., op. cit., pp. 72–78.

$$\begin{aligned} H_{\theta(\text{total})} = {} & \frac{I_0}{2\pi\rho} \sin \omega(t - r_0/c) \\ & + \frac{I_0}{2\pi\rho} \sum_{k=1}^{\infty} \Big\{ \sin [(r_{-(2k-1)} + l - r_{-2k})\omega/2c] \\ & \cdot \cos [\omega t - (r_{-(2k-1)} + l + r_{-2k})\omega/2c] \\ & + \sin [(r_{(2k-1)} + l - r_{2k})\omega/2c] \\ & \cdot \cos [\omega t - (r_{(2k-1)} + l + r_{2k})\omega/2c] \Big\}. \end{aligned} \tag{17}$$

The subscripts associated with the r's are defined in Fig. 5(a).

Eq. (17) represents a generalization of (1) and reduces to (1) for small ω. Eq. (17) may now be used to estimate the error associated with the quasi-stationary equations obtained in the previous section.

Closer inspection of (17) reveals that the series is absolutely convergent for all finite values of ω. Also, the coefficients of the succeeding terms are monotonically decreasing positive quantities in all cases for which $(r_{\pm 1} + l - r_{\pm 2})\omega/2c$ is less than $\pi/2$. This can be seen physically from Fig. 5(a); $(r \pm (2k-1) + l - r \pm 2k)$ decreases steadily for increasing k, ultimately approaching zero as k approaches infinity. Since the first term of (17) corresponds to the low frequency approximation, the $k=1$ term may be taken as the first-order high-frequency correction term. Thus:

$$\begin{aligned} H_{HF} \cong {} & \frac{I_0}{2\pi\rho} \{ \sin \omega(t - r_0/c) + \sin [(r_{-1} + l - r_{-2})\omega/2c] \\ & \cdot \cos [\omega t - (r_{-1} + l + r_{-2})\omega/2c] \\ & + \sin [(r_1 + l - r_2)\omega/2c] \\ & \cdot \cos [\omega t - (r_1 + l + r_2)\omega/2c] \}. \end{aligned} \tag{18}$$

H_{HF} is used in place of H_θ to refer to the approximate field of the conductor and its image in the end walls at high frequencies.

The total field inside the enclosure is predominantly due to the wire and its first image in the ceiling. It was shown that for particular cases the wire and its nearest image contribute more than 80 per cent of the total field computed using (1). Thus if results of (1) and (18) applied to these two conductors only are compared, a fairly good measure of the error should be obtained. Let the high frequency field due to the wire itself be H_{HF1}, and the high frequency field due to the image in the ceiling be H_{HF2}. Also, let the corresponding low frequency field be H_{LF1}, and H_{LF2}. Then

$$\frac{|H_{HF1} + H_{HF2}|}{|H_{LF1} + H_{LF2}|} - 1 = \epsilon. \tag{19}$$

ϵ is the approximate error to be expected when the low frequency equations are used.

A specific example, chosen to simulate a practical condition, will be used to illustrate the errors that may be expected in the high frequency range. Let $l = 5.15$

meters, $\rho_1 = 1.0$ meters, and $\rho_2 = 1.61$ meters. The subscript 1 indicates a distance from the wire; the subscript 2 indicates a distance from the image. The point at which the field is measured is one meter from the receiving end of the wire. Table I below gives the ratio ϵ computed at several frequencies for the conditions stated.

TABLE I

Freq	ϵ	ϵ_a
5 mc	−0.5 per cent	−1.2 per cent
10	−1.6 per cent	−2.5 per cent
15	−3.0 per cent	−4.1 per cent
20	−4.2 per cent	−5.5 per cent
25	−5.0 per cent	−7.0 per cent

A somewhat coarser estimate of the error will be obtained if the phases of the terms in (18) are ignored. The column labeled ϵ_a in Table I was computed on this basis. ϵ_a gives a more pessimistic result and may be viewed as a safer estimate.

Validity of the Method of Images at Very Low Frequencies

An important assumption inherent in the use of the method of images is that the walls of the enclosure are perfectly conducting. In the case of real conductors, there is a question as to how wall conductivity affects the validity of the image method.

In order for the image method to apply rigorously, it is required that the current flow in the wall be all at the surface. For slightly imperfect conductors the current in the wall flows primarily at the surface but, nevertheless, extends somewhat into the conductor. The depth of penetration is known to be an inverse function of frequency. Thus, as the frequency is reduced, the image method can be expected to give less accurate results. Most work using loop antennas does not involve frequencies below about 15 kc. On the assumption that the wall is sufficiently thick so that the current density is nearly zero at the far boundary of the wall, no appreciable error arises when the image method is used above this frequency. For this condition the fields inside the conductor become nearly zero at the far boundary and, hence are nearly zero outside this boundary.

In the event that a thin wall is used so that an appreciable current density exists at the far boundary, the fields inside the wall do not fall to zero. In this case a method for estimating the magnitude of error involved in using images is useful.

Refer to Fig. 6 in which a long current-carrying wire is shown near a conducting wall. The thickness of the wall and the wire diameter will be assumed small compared to any other dimensions. Assume that the field at point P is to be found. The image of point P on the far side of the wall is shown at point P'. Suppose the magnetic field at point P is given by H_T and the field at point P' is kH_T. The field at P and P' can be computed by superposition of the fields due to the current in the wire alone and the current in the wall alone. Let the field due to the current in the wire be H_1 and H_1' at P and P' respectively. Since P and P' are symmetrical with respect to the wall then the fields due to the wall currents alone at these two points are equal and opposite. Call this field H_2 at P and $-H_2$ at P'. Thus

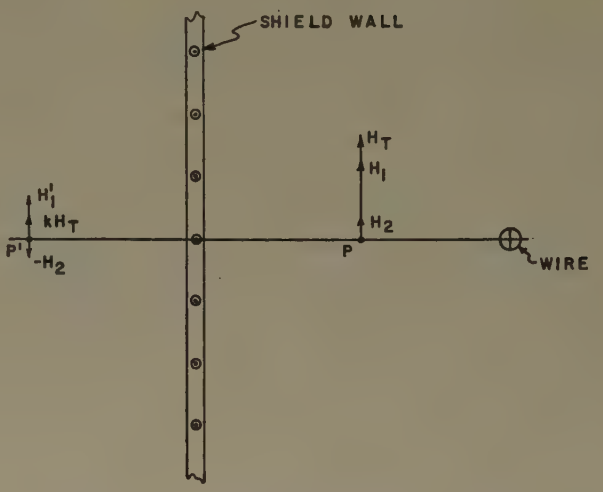

Fig. 6—Superposition of fields due to current in the wire and current in the shield wall.

$$H_1' - H_2 = kH_T$$

and

$$H_1 + H_2 = H_T.$$

Thus

$$H_T = \frac{H_1 + H_1'}{1 + k}. \tag{20}$$

Note that for $k = 0$, $H_T = H_1 + H_1'$ is simply the field at P due to the wire and its image in the wall. Thus the factor, k, may be considered the error in the field at P due to imperfect wall conductivity. Although k may be complex it is sufficient here to take its absolute value. It may be inferred generally that the image method is valid where the conducting enclosure is a good shield. If there is any doubt the value of k can be determined by relative field measurements at two symmetrical points as indicated in the figure. Although the arguments above were based on a single wall of very large area, the conclusions may be extended to the case of a practical enclosure without introducing excessive uncertainty in the estimate of the error. It should be pointed out that this pertains to single shield enclosures; in the case of double shield enclosures the shielding ability of the inner wall must be determined.

Conclusion

In light of the foregoing the following remarks can be made regarding the "transmission line method." In the low and medium frequency range the field under the wire can be determined with an accuracy better than

1 per cent when (8) or (9) is used. It is recommended that the ratio of height to width of the enclosure be in the order of one, or greater, to ensure this accuracy for any reasonable choice of the other parameters. Errors in measurement are not included in this figure. Where any question arises regarding the error involved the bounds, (6) and (7) may be used. In the high frequency range the accuracy obtained using (8) or (9) can be expected to fall off in some installations, such as the one described. If the standing waves can be avoided on the line an estimate of the error can be obtained using (19). In the very low frequency range the quality of the enclosure as a shield determines the validity of (8) and (9). By making a relative measurement of the field at two points, and by the use of (20), an estimate of the error in this range can be obtained.

ACKNOWLEDGMENT

The author wishes to thank Messrs. E. I. Hawthorne, F. T. Nicholson, and B. Epstein, for their invaluable suggestions and assistance. This work was done under Contract NObsr 49128 with the Bureau of Ships.

Correspondence

Comment on "Instantaneous Frequency"*

R. W. Johnson's note regarding instantaneous frequency in the correspondence columns of the PROCEEDINGS OF THE I.R.E.[1] is noted.

With the plethora of comments about instantaneous frequency which had been appearing in these columns, and noting a discussion of the subject in the *Proceedings of Commission VI of URSI*, at the Xth General Assembly in Sydney, Australia, in August, 1952, I wrote to Dr. B. van der Pol, President of that Commission, in December, 1953, calling his attention to the discussion in these columns. His following reply of February, 1954 leaves no question about his position in the matter.

"Thank you for your letter of December 15th. It was very kind of you to draw my attention to the letters published in the correspondence columns of the PROCEEDINGS OF THE I.R.E.; they had escaped my notice.

"I have now gone through them and noted that they concern the idea of 'Instantaneous Frequency,' which I introduced in radio some 25 years ago in PROC. I.R.E., vol. 18, p. 1194; 1930, of which I am sending you a reprint under separate cover.

"I am fully confident that my definition was right, proper and useful, and still is."

MILLETT G. MORGAN
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* Received by the IRE, August 4, 1954.
[1] Vol. 42, p. 1024; June, 1954.

Comment on "Properties of Some Wide-Band Phase-Splitting Network"*

D. G. C. Luck[1]

J. S. LeGrand of our Laboratories recently made use of some RC 90-degree-phase-difference networks of the type shown in Fig. 2(c) and Fig. 2(e) of Dr. Luck's very useful article. When the design equations given in Figs. 2(c) and 2(e) were used, unsatisfactory circuit performance resulted.

To double-check the design equations, I have independently analyzed the two circuits involved and find that the improper performance was a result of the following errors in Dr. Luck's design equations.

In Fig. 2(c) the equation for both attenuation and required magnitude of the 180°-phase-shifted generator should be:

$$k = \frac{1}{2}\,\frac{(1-2Q)}{(1+2Q)}. \tag{1}$$

In Fig. 2(e) the number two (2) is missing from six of the equations and I will rewrite the corrected equations below:

$$Q = \frac{1}{2}\,\frac{1}{(R_3/R_2+1)}. \tag{2}$$

$$k = \frac{1}{2}\left(\frac{1-2Q}{1+2Q}\right). \tag{3}$$

$$R_2 = 2R_3/[1/Q-2]. \tag{4}$$

$$R_1 = 2R_3/[1/Q^2-4]. \tag{5}$$

$$C_2 = \tfrac{1}{2}[1/Q-2]C_3. \tag{6}$$

$$C_1 = \tfrac{1}{2}[1/Q^2-4]C_3. \tag{7}$$

* Received by the IRE, July 21, 1954.
[1] PROC. I.R.E., vol. 37, pp. 147–151; Feb. 1949.

Incidentally, it may be helpful to other readers to point out the fact that the design parameters f_1, f_2 and Q of Dr. Luck's useful article can be obtained as follows:

1. Obtain the Smithsonian Elliptic Function Tables by C. W. and R. M. Spenceley.
2. From your required f_{max} and f_{min} calculate

(a) $$\theta = \sin^{-1}\sqrt{1-(f_{min}/f_{max})^2}.$$

3. If θ is not an exact integral number of degrees modify (f_{min}/f_{max}) slightly so as to make θ be an integer and still satisfy your requirement. (This procedure removes need for double interpolation in above tables.)
4. Using the above tables calculate

(b) $$a_1 = \frac{cn_\theta(K\theta/8)}{sn_\theta(K\theta/8)}.$$

(c) $$a_3 = \frac{cn_\theta(5K\theta/8)}{sn_\theta(5K\theta/8)}.$$

5. Finally obtain f_1, f_2 and Q from

(d) $$\frac{f_2}{f_{max}} = \frac{f_{min}}{f_1} = \sqrt{a_1 a_3}.$$

(e) $$\frac{1}{Q} = \left(\sqrt{\frac{a_1}{a_3}} + \sqrt{\frac{a_3}{a_1}}\right).$$

The background material for the above procedure is contained in the basic articles by Darlington[2] and Orchard.[3]

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[2] S. Darlington, "Realization of a constant phase difference," *Bell Sys. Tech. Jour.*, vol. 29, pp. 94–104; January, 1950.
[3] H. J. Orchard, "Synthesis of wideband two-phase network," *Wireless Eng.*, pp. 72–81; March, 1950.

Correspondence

A Note on the Fourth Dimension*

It may be time that the definition of time be re-examined. The arbitrary intervals set up as standards may have acted as a bottleneck in progress to fully describe the physical environment.

Consider that the passage of time is a function of the change of state of the matter being described. This hypothesis broadens the accepted concept to include nonuniform time changes in general (and in particular, only an approximation of perfect uniformity). The entropy of an ensemble is a measure of its exact organization state. Any change in this entropy is therefore a passage of time.

This entropy is a function of the system internal probabilities so that when they remain unchanged (within the accuracy of measure), then no time has passed for the system. This implies that time is not the fourth dimension and that relative entropy is. (Relative entropy is used to define the static time of the system considered in relation to its final state and thereby it will indicate "age" toward "death" or "uselessness.") The comparison of "times" of different objects may be readily covered by considering each in relation to the entire containing system.

It is considered possible to have negative change of "time" or a decrease in entropy for a particular subject when energy with information is being supplied and the subject is taken to be the entire system in question. The inevitable increase in universal entropy is reflected in the fact that "time stands still for nothing." Yet as far as humans can measure, things of constant probability and entropy are timeless, such as a coin flip or a rock.

A consideration of dimensional analysis should be met early in the proposed concept of relative entropy as the fourth dimension. Since both entropy and relative entropy are both dimensionless quantities and the first three dimensions are measured in arbitrary units, the dissimilarity must be overcome. Two methods are available. Either correct relative entropy to a proper length unit of measure by multiplying by some absolute constant (such as was accomplished in multiplying time by the velocity of light) or reduce the first three dimensions to unitless quantities. Since no absolute length seems defined by nature, the first possibility appears impractical (as well as arbitrary). The second, however, seems made to order. If the system bounds are chosen before attempting to write number values of description, then relative terms may be used throughout. The first three dimensions would be a measure of the ratio of the distance from the origin to the position described, to the distance to the boundary of the system in each direction. The fourth dimension would similarly be a measure of the ratio of the entropy being considered to the maximum entropy of the system.

The quantities thus far considered are the static dimensional values. Since that which is usually of concern is the change in dimensions, the dynamic values must be examined. Just as $\Delta l = l_2 - l_1$ and $\Delta t = t_2 - t_1$ so it should be defined that $\Delta h = h_2 - h_1$ where h represents the relative entropy H/H max. Next, one unnecessary restriction previously incumbent seems to have been overcome: the instantaneous rate of time change was uniform; however, the instantaneous rate of relative entropy change (the relative entropy gradient) is nonuniform and therefore more explicit.

The special case of periodic variation of relative entropy may be of interest. Consider an amplitude varying information containing signal. Fourier Analysis will break it down into sinusoidal variations in relative entropy (in the entropy domain).

A rather unusual shaped curve would appear (Fig. 1) upon plotting a single sinusoidal component at some constant "frequency" (angular rate of vector rotation ω of x_r). This is due to the nonuniformity of h. The sinusoidal waveshape could be preserved if ω were such a function of h that the product ωh would be linear.

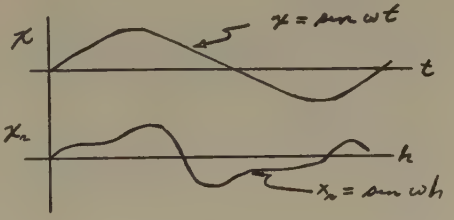

Fig. 1

It might also be profitable to find the relation between h and t. This would describe the real nonuniform time in terms of the fictitious uniform old definition of time.

Since physical entropy is a direct measure of information, the specification of any point in accordance with the proposed system falls into values of relative position in space and relative information. This abides with the complete universal specification of any system in terms of material, energy and information (with no time measure). It should also be pointed out that since R (the redundancy) $= 1 - (H/H \text{ max}) = 1 - h$, then the redundancy is also a measure of the true "time" of the system. It is also important that the quantized states of energy enforces the concept of discontinuous "time" change together with all its implications. It is suggested that natural biological activity might be more readily specified as variations in entropy rather than cycles in arbitrarily uniform time which is astronomical and completely unrelated to the living organism.

After this concept had been set down it was found that some related ideas were discussed in the literature. In particular A. S. Eddington, in his "The Nature of the Physical World," stated the following:

"Having convinced ourselves that the two things are connected, (time and entropy-gradient) we must conclude that there is something as yet ungrasped behind the notion of entropy—some mystic interpretation, if you like—which is not apparent in the definition by which we introduced it into physics. In short we strive to see that entropy-gradient may really be the moving on of time (instead of vice versa). (p. 95)

"Entropy-gradient is then the direct equivalent of the time of consciousness (a mental time concept) in both its aspects. Duration measured by physical clocks (time-like interval) is only remotely connected. (p. 101.)

"Temporal relation and entropy-gradient, both rigorously defined in physics, are entirely distinct and in general are not numerically related." (p. 102)

It is felt that there remains considerable difference between the implications of these statements and the concept proposed in this paper.

This note is presented to indicate a line of reasoning which may or may not prove fruitful. It should be considered in the light of relativity and information theory as well as the basic laws of classical mechanics before final evaluation may be performed.

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TABLE I
Parameters and Symbols

	Standard	Proposed
Relative position to origin	x, y, z	$\frac{x}{x_{max}} = x_r, y_r, z_r$
Change of position	$\Delta x, \Delta y, \Delta z$	$x_{r2} - x_{r1} = \Delta x_r, \Delta y_r, \Delta_r$
Time (absolute)		H of the system
Relative time		$\frac{H}{H_{max}} = h = g(t).$
Gradient time	$\frac{dx}{dt} = \text{const.}$	$\frac{dx_r}{dh} = \text{variable}$
Frequency	$f = \frac{\sin^{-1} x}{2\pi t}$	$f = \frac{\sin^{-1} x_r}{2\pi h}$

* Received by the IRE, July 22, 1954.

Correspondence

High Latitude Ionospheric Observations Using Extra-Terrestrial Radio Waves*

The scintillation of localized extra-terrestrial radio sources[1,2,3] and the absorption of the diffuse background of extra-terrestrial radio waves[4] have been used by various workers to investigate the ionosphere at temperate latitudes. This letter describes some preliminary observations of the effect of the ionosphere upon the incoming radio waves at College, Alaska, situated approximately 275 kms. south of the auroral maximal zone as defined by Vestine.

The apparatus used in this investigation consisted of a stable, high-gain receiver and a directional antenna, the output of the receiver being recorded continuously by a pen recording instrument. The antenna, a twin Yagi array with beam widths between half power points of approximately 26 degrees and 50 degrees in the two principal planes, was directed at azimuth 215 degrees east of true North and elevation 70 degrees, the beam being oriented so that it was narrow in azimuth and broad in elevation.

Owing to the rotation of the earth, the beam swept out each day a broad strip of the sky which included the localized sources in Cassiopeia and Cygnus (declinations +58 degrees 32′ and +40 degrees 35′ respectively). The stability of the receiver gain was monitored by switching the receiver from the antenna into a dummy antenna every half hour for a period of 20 seconds—in this way, any drifts due to changes in the gain of the receivers (drifts which might otherwise be attributed to changes in the received antenna power) were revealed by corresponding drifts in the positions of these periodic reference levels.

The records obtained in this way over a two-month period commencing March 16, 1954, have been inspected for two main effects: (a) absorption of the diffuse background of extra-terrestrial radio waves and (b) scintillation of the intense localized radio sources in Cassiopeia and Cygnus.

Absorption

In the absence of ionospheric effects, the intensity of the extra-terrestrial radio waves received by a fixed antenna at a given frequency is a function only of sidereal time. The degree of absorption of the diffuse background of extra-terrestrial radio waves may therefore be determined from the ratio of the signal actually received to the signal received at that sidereal time under conditions of negligible ionospheric absorption. The curve showing the sidereal variation of extra-terrestrial radio power for this equipment is shown in Fig. 1 and was obtained by averaging seven records taken on days when ionospheric sounding equipment, and equipments monitoring distant radio stations, showed that ionospheric absorption was unusually low. The scatter between the records on quiet days was small (rms deviation of signal power at a given sidereal time less than 2 per cent); it is considered that any residual absorption of 65 mc signal on these days was less than 1 per cent.

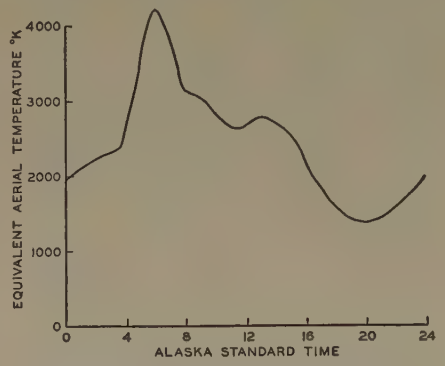

Fig. 1—Diurnal variation of received extra-terrestrial radio power. (The time marks are those for May 10, 1954).

The degree of absorption was measured every half hour throughout the two-month period by comparing each day's record in turn with the standard no-absorption record obtained as above.

The frequency of occurrence of strong absorption at 65 mc (absorption in excess of 10 per cent) is plotted in Fig. 2 as a function of the time of day. It will be seen that such absorption is most common at midday, and occurs least frequently during the late evening. This 65 mc absorption was found to be strongly correlated with no-echo conditions (polar blackouts) on the C-3 Ionospheric Sounding Equipment, the two phenomena showing a similar diurnal variation (Fig. 2), and similar day-to-day variations. (Fig. 3).

In Fig. 4, the daily sum of the three-hour Geomagnetic K indexes for College, Alaska, are compared with the number of half-hourly 65 mc observations per day showing absorption greater than 10 per cent. The correlation coefficient between the two phenomena was 0.48, indicating a strong link between ionospheric absorption and the variability of the earth's magnetic field.

Scintillation

The scintillation work described here, although limited to two months' observations at high angles of elevation, is sufficient to indicate that marked differences exist between the scintillation phenomena in England and Alaska. Thus, scintillations are more common in Alaska than in England and apparently do not show the same diurnal variation of activity. (In England, zenithal fluctuations are found to be a nighttime phenomenon and are rarely observed after 0800 G.M.T.; in Alaska, the fluctuations occurred with equal frequency during the four hours preceding, as in the four hours immediately following, 0800 Alaska Standard Time.) Another difference was the variability of the high latitude fluctuations, whose characteristics usually changed considerably during a three-hour transit period; on several occasions the transit of a source was undisturbed except for one (or occasionally two) short bursts of fluctuations.

One point of considerable interest is that scintillations were always observed whenever polar blackouts or marked 65 mc absorption occurred during the transits of the localized radio sources through the aerial beam. Since "radio star" scintillations are caused by the presence of localized irregularities in the electron density of the ionospheric layers, this must mean that the lateral distribution of ionization during periods of absorption is nonuniform. It should therefore be possible, by applying existing techniques to the study of these scintillations, to obtain information about the size, shape, stability, and movements of the

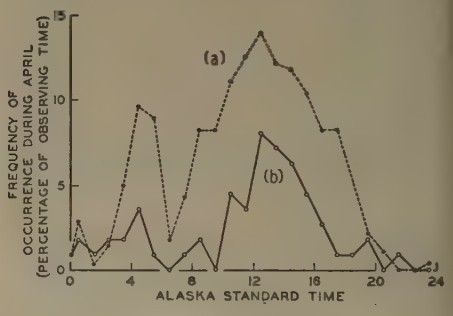

Fig. 2—Diurnal variation of (a) occurrence of polar blackouts and (b) of strong (i.e. greater than 10 per cent) absorption of extra-terrestrial radio waves.

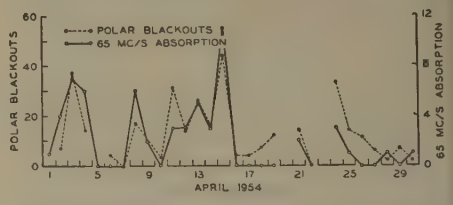

Fig. 3—Daily frequency of detection of polar blackouts (120 observations per day) and strong 65 mc absorption (48 observations per day) during April, 1954.

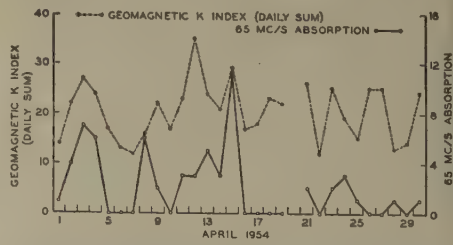

Fig. 4—Comparison of daily sum of three-hour Geomagnetic *K* indexes for College, Alaska, and number of half-hourly 65 mc observations per day showing absorption greater than 10 per cent.

* Received by the IRE, June 18, 1954. This work has in part been financed by the National Science Foundation.

[1] C. G. Little and A. Maxwell, "Fluctuations in the intensity of radio waves from galactic sources," *Phil. Mag.*, vol. 42, pp. 267–278; March, 1951.

[2] A. Hewish, "The diffraction of galactic radio waves as a method of investigating the irregular structure of the ionosphere," *Proc. Roy. Soc., A*, vol. 214, pp. 494–514; October 9, 1952.

[3] A. Maxwell and C. G. Little, "A radio-astronomical investigation of winds in the upper atmosphere," *Nature*, vol. 169, pp. 746–747; May 3, 1952.

[4] C. A. Shain, "Galactic radiation at 18.3 mc/s," *Aust. Jour. Sci. Res.*, vol. 4A, pp. 258–267; September, 1951.

Correspondence

ionospheric irregularities, during periods of heavy absorption when ionospheric echo-sounding techniques are unusable owing to the heavy attenuation of the reflected wave.

It is proposed to extend the present range of the experiments in order to investigate more fully the absorption and scintillation of extra-terrestrial radio waves. The use of these extra-terrestrial radio waves for ionospheric studies at high latitudes appears to have many advantages over normal echo-sounding techniques, e.g., the absence of the need to build and operate expensive radio transmitters; the use of higher frequencies enabling directional antennas to be built and hence permitting the investigation of a wide area of the ionosphere from a single station; the simplification of the results owing to the small effect of the earth's magnetic field on radio wave propagation at frequencies well above the gyro-frequency; the possibility of obtaining information about regions of the ionosphere above the level of maximum electron density; and the opportunity of studying ionospheric irregularities at times when frequencies below the critical frequency are heavily absorbed.

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Power Flow and Equivalent Circuits of Traveling-Wave Tubes*

C. C. Wang† to J. R. Pierce‡, September 30, 1953

In this equivalent circuit of traveling-wave tubes, using the notation (Fig. 1) in your book,[1]
but, adding losses to the line, we have

$$\frac{\partial i}{\partial z}+\frac{\partial I}{\partial z}=-V(j\omega C+G) \qquad (1)$$

$$\frac{\partial V}{\partial z}=-I(j\omega L+R). \qquad (2)$$

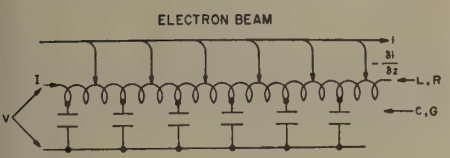

Fig. 1

If we take the power products

$$(1)^*V+(1)V^*+(2)^*I+(2)I^*$$

then we have

$$\frac{1}{4}\left\{\frac{\partial(VI^*+V^*I)}{\partial z}+\left(V\frac{\partial i^*}{\partial z}+V^*\frac{\partial i}{\partial z}\right)\right\}=-\frac{1}{2}(I^2R+V^2G).$$

Since the first term represents the variation of power flow in the circuit, and the last term represents the power dissipation in the line, it is evident that the product,

$$-\frac{1}{4}\left[V\left(\frac{\partial i^*}{\partial z}\right)+V^*\left(\frac{\partial i}{\partial z}\right)\right]$$

must represent the power fed into the line by the beam.

On the other hand, the rate of change of kinetic energy of the beam with respect to the distance must be the power product of the electric field $\partial V/\partial z$ and the current i and this amounts to

$$\left[\left(\frac{\partial V^*}{\partial z}i+\frac{\partial V}{\partial z}i^*\right]\frac{1}{4}\right.\cdot$$

Thus we face the problem of nonequality of these two energy expressions. The discrepancy just amounts to

$$\frac{1}{4}\frac{\partial(V^*i+Vi^*)}{\partial z}$$

which is the algebraic sum of the two. The question is whether we should consider the product $\frac{1}{4}(V^*i+Vi^*)$ as a sort of energy flow in order to keep the energy balance sheet.

Fig. 2

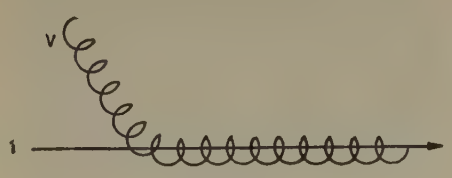

Fig. 3

This energy flow exists as long as V and i are coupled. If so, what happens to this energy when V and i are suddenly decoupled such as at the end of the tube (Fig. 2)? Or, if a current is sudddnly introduced at the beginning of the circuit with a voltage V (Fig. 3)?

According to the equivalent circuit at these junctions there are large $\partial i/\partial z$ introduced at the circuit and large electric field variations $\partial V/\partial z$ at the beam, but presumably they are finite, integrable within the transition section. Therefore, at these junctions, you would expect some energy interchange between the beam and the circuit.

On the other hand, if $\frac{1}{4}(V^*i+Vi^*)$ cannot be considered as energy flow, then something must be done to restore the energy balance. This requires an extra condition to be imposed upon the traveling-wave tube equation.

I might add that if this is a discrepancy at all, it only amounts to an order of C times the normal circuit energy. Therefore, it will be negligible anyway, if we assume C is small and so on. However, in the so-called large C cases, then this is just the order of magnitude we are dealing with in making our calculations.

J. R. Pierce to C. C. Wang, October 4, 1953

I think we should regard P_e

$$P_e=\tfrac{1}{4}(V(I+i)^*+V^*(I+i))$$

as giving the total electromagnetic power flow, as you suggest, so that for a lossless circuit our equation is

$$\left(\frac{\partial P_e}{\partial z}-\left(i^*\frac{\partial V}{\partial z}+i\frac{\partial V^*}{\partial z}\right)\right)=0.$$

Suppose we have, as shown in Fig. 1, a helix with a wire current I and a beam current i, ending in a tube at helix voltage into which the beam current i flows without being acted on by a longitudinal field.

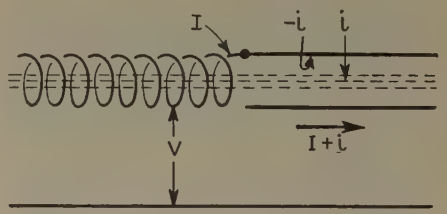

Fig. 1

Inside the tube there will be a beam current i and a wall current $-i$. On the outside of the tube there will be a current $I+i$. Hence, the circuit power delivered by the current flowing on the tube against the voltage V will be

$$\tfrac{1}{4}(V(I+i)^*+V^*(I+i))$$

just as we have assumed. The integral of

$$\frac{1}{4}\left(i^*\frac{\partial V}{\partial z}+i\frac{\partial V^*}{\partial z}\right)$$

from the beam source to the tube will be left in the beam as kinetic energy.

C. C. Wang to J. R. Pierce, November 13, 1953

Suppose we have a waveguide in which a free TM mode exists and it has a propagation constant γ_m, such that $\beta_m^2=\gamma_m^2-\gamma_0^2$ where $\gamma_0^2=-\omega^2\mu\epsilon$, $\gamma_0=\sqrt{\mu\epsilon}$ and if E_{zm} is the longitudinal component of the electric field, it will satisfy the equation

$$\nabla^2E_{zm}+\beta_m^2E_{zm}=0$$

where ∇ represents the nabla operator in the transverse direction only. The transverse electric field will be proportional to ∇E_{zm} while the transverse magnetic field will be proportional to $\vec{k}\times\nabla E_{zm}$, where $\vec{k}$ is a unit vector in the z direction. Let us assume any electric field and magnetic field can be ex-

* Received by the IRE, August 17, 1954.
† Sperry Gyroscope Co., Great Neck, L. I., N. Y.
‡ Bell Telephone Labs., Inc., Murray Hill, N. J.
[1] J. R. Pierce, "Traveling-Wave Tubes," D. Van Nostrand Co., New York, N. Y., p. 9; 1950.

Correspondence

panded in a series of modes which read like

$$\vec{E} = -A_m \frac{\gamma_m}{\beta_m^2} \nabla E_{zm} + \vec{k} B_m' E_{zm} + \cdots \quad (1)$$

$$\vec{H} = -B_m \frac{\gamma_0}{\beta_m^2 Z_0} \vec{k} \times \nabla E_{zm} + \cdots \quad (2)$$

where we use the contraction of the index m to indicate the summation of all the modes and $+\cdots$ indicates that the series is not complete, e.g., there are TE modes present. However, for this discussion, we only have to pay attention to one TM mode at a time. $Z_0 = \sqrt{\mu/\epsilon}$.

A_m, B_m, B_m' are three arbitrary functions of z only while E_{zm} is understood to be functions of the transverse space co-ordinates only.

For simplicity, let us assume that we only have a longitudinal current in the traveling-wave tube so that the current density i_d can be expanded as a series of

$$i_d = \vec{k} J_m E_{zm} + \cdots \quad (3)$$

J_m is again a function of z only.

If we try to substitute these into the Maxwell's curl equations, we must have

$$B_m' = B_m - \frac{Z_0}{\gamma_0} J_m \quad (4)$$

$$\frac{dB_m}{dz} + \gamma_m A_m = 0 \quad (5)$$

$$\frac{dA_m}{dz} + \gamma_m B_m = \frac{\beta_m^2}{\gamma_m} J_m \frac{Z_0}{\gamma_0}. \quad (6)$$

[1]If we further limit our interest to a filament of current at a particular location of the waveguide, so that the current i is i_d multiplied by a small area a and at this point E_z assumes a particular value of E_{zm0}, thus, by (3) we have $\vec{i_d} = \vec{k} i/a$.

$$J_m = \frac{\iint \frac{i}{a} E_{zm} da}{\iint E_{zm} E_{zm} da} = \frac{i E_{zm0}}{\iint E_{zm}^2 da}. \quad (7)$$

Define also an impedance Z such that

$$Z = \frac{\beta_m^2}{\gamma_m^3} \frac{Z_0}{\gamma_0} \frac{E_{zm0}^2}{\iint E_{zm} E_{zm} da}. \quad (8)$$

Let us also renormalize A and B into voltage V and current I, such that

$$V = \frac{E_{zm0}}{\gamma_m} A_m$$

$$I = \frac{E_{zm0}}{\gamma_m Z} B_m. \quad (9)$$

The reorganized differential equations then take the following forms:

$$\frac{\partial V}{dz} + \gamma_m Z I = \gamma_m Z i \quad (10)$$

$$\frac{\partial I}{\partial z} + \frac{\gamma_m}{Z} V = 0. \quad (11)$$

Furthermore, the axial electric field at the beam

$$E_z \text{ at beam} = B_m' E_{zm0}$$

$$= \gamma_m Z I - \left[\frac{E_{zm0}^2 \frac{Z_0}{\gamma_0}}{\iint E_{zm} E_{zm} da} \right] i. \quad (12)$$

The above three equations suggest the equivalent circuit for a traveling-wave tube, as in Fig. 1. According to this, the beam and the circuits are coupled together through a distributed mutual impedance $\gamma_m Z$ equal to the distributed impedance of the line itself. Also, the beam appears to have a self-

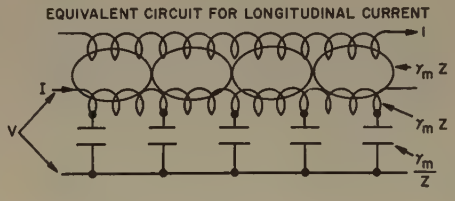

Fig. 1

impedance which is the bracket term in (12). The electric field induced by a current i into the line is $i\gamma_m Z$ while the electric field due to the line current I at the beam is $I\gamma_m Z$. The self-induced field at the beam due to its own current is reactive in nature since Z_0/γ_0 is essentially imaginary for nonlossy dielectrics. This self-induced field is in the nature of a space charge force. I believe since the space charge forces are nonsolenoidal, the space charge force represented by the second term of (12), which is an expansion in terms of a solenoidal mode alone, will not be complete. We may add an arbitrary nonsolenoidal and irrotational field without affecting the above equivalent circuit. I suppose in the end it can all be lumped into one single term proportional to the square of the plasma frequency as it was usually done. However, much remains to be done to clear the point.

As to the energy interchange, it is simply the time average of $\gamma_m Z i$ and I. The electromagnetic power flow will be represented by the time average of VI. This checks with the Poynting Vector product of (1) and (2) provided that we assume all the energy is contained in the TM mode. I suppose if a mixed mode is present instead of (8), the impedance Z can be redefined by using the actual value of the Poynting Vector including the TE component. The conservation of energy is seen to be preserved when we take proper products of (10), (11), and (12).

Now, if we make a further transformation

$$I = I' + \quad (13)$$

then (10) and (11) become

$$\frac{\partial V}{\partial z} + (\gamma_m Z) I' = 0 \quad (14)$$

$$\frac{\partial I'}{\partial z} + \frac{(\gamma_m) V}{Z} = -\frac{\partial i}{\partial z}. \quad (15)$$

This immediately puts in a form equivalent to your original circuit. The fact that this transformation leaves the second order wave equation in V unchanged makes it appear that (14) and (15) can also serve as the basic equivalent circuit equations. However, this is so provided that we have to recognize the electromagnetic power as $V(I'+i) = VI$ instead of VI', as we discussed before. I' can not then be considered as the complete circuit current.

In conclusion, the above derivation is basically the same as your original derivation of normal mode equation except that the coefficients are left in the equations and two basic transmission line (10) and (11) are directly obtained, instead of one single wave equation. These equations suggest that we should use another form of equivalent circuit so that we can truthfully say that the circuit voltage-current product represents all the electromagnetic power flow there is.

This derivation also shows that the circuit voltage V is simply a measure of the transverse electric field and the circuit current I is a measure of the transverse magnetic field. I have the difficulty of trying to relate the equivalent circuit current to any actual localized current flow along metallic conductors. It is easy to do in the case of a TEM mode of coaxial transmission, since this is undoubtedly the same as the current actually flowing along the conductors.

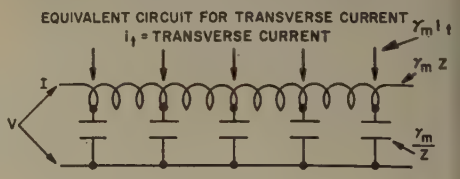

Fig. 2

However, in the case of a mixed mode like that of a helix, we can not fix a particular point where we can make a measure of this equivalent circuit current. Certainly it is not the current flowing along the helix wire. Therefore, to relate this current to actual beam current by using the divergence equation and the equation of continuity of currents (displacement current and conduction current), will need further clarification.[2]

Adding to the above derivation, when transverse currents are present, we may draw the equivalent circuit as in Fig. 2.

[2] Pierce, ibid.

P. K. Tien† to C. C. Wang, March 9, 1954

One thing which bothers me is as follows: We always have a conception that the coupling between the beam and the circuit is

† Bell Telephone Labs., Inc., Murray Hill, N. J.

[1] The following derivation is strictly true only for nonlossy waveguides so that γ_m and β_m are pure imaginary and Z is real. For the lossy case, other definitions of current, voltage, and impedance must be used and the equivalent circuit will be changed accordingly.

Correspondence

capacitive, and the beam generally acts as an inductive element to the axially directed electric field. In your equivalent circuit, however, the coupling turns out to be inductive.

A study of your (10), (11) and (12) indicates that in order for these equations to be consistent with your equivalent circuit, the mutual impedance is $\gamma_m z$ which is inductive and

$$\frac{\partial V}{\partial z} + \gamma_m z I + \gamma_m z(-i) = 0. \qquad (10)$$

The use of $-i$ instead of i in the above equation puzzles me, although as the equations stand, it seems straightforward that the $-i$

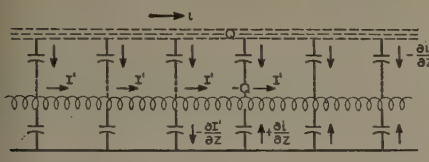

Fig. 1

of the electron beam acted with the mutual impedance $\gamma_m z$, enters the transmission line equation (10) and so in your equivalent circuit. It may however confuse the beginner why $-i$ should be used instead of i.

To explain this, we say that the coupling is actually capacitive. Let's draw a physical picture. Assuming that the coupling is capacitive, any positive charge in the beam will induce a negative charge on the circuit. When the beam current varies, the charge on the circuit varies accordingly and this is accomplished by an additional current flowing on the shunt element of the circuit in a direction from ground to the circuit, as illustrated in Fig. 1, which of course is the Pierce original equivalent circuit. As mentioned by Pierce, without the electron stream, the electromagnetic power is VI'^*. Because of the electron stream and the capacitive coupling, the total current flow in the shunt element is $\partial(I'+i)/\partial z$ and so the power is $V(I'+i)^*$.

Your equivalent circuit traces out very clearly the relation between the power-flows, except that the use of $-i$ and the inductive coupling may mislead with regard to the actual physical concept.

We should note that in your first letter I is the circuit current as defined in Pierce's book while in your second letter I' is the same current. Our notation follows that in your letters.

A Rapid Method for Matching High Frequency Transmission Systems*

Recent developments in the manufacture of dielectric materials make the following a simple and effective method for rapid matching in waveguide and other high frequency transmission systems. A slug of dielectric material having the same cross section as the guiding system, and a quarter wavelength long (in the dielectric) is inserted into the guide. The dielectric constant is chosen so that the slug produces a vswr equal to the mismatch which is to be eliminated. The slug is then moved along the guide until a match is achieved.

The vswr produced by such a dielectric slug may be calculated either by requiring continuity of tangential E and H across the two surfaces of the dielectric, or by a consideration of the wave impedances normal to the surfaces. The result of either calculation is that the dielectric constant required to produce a given standing wave ratio "s" is given by

$$\epsilon = s\left[1 - \left(\frac{\lambda}{\lambda_c}\right)^2\right] + \left(\frac{\lambda}{\lambda_c}\right)^2$$

where λ_c is the free space cut-off wavelength of the guiding system, λ is the free space operating wavelength, and ϵ is the relative dielectric constant of the slug. It is to be noted that this relation is fairly general; it holds for all TE waves in any uniform cylindrical guiding system, i.e., systems which can be generated by translating a cross-section parallel to a fixed straight line.

It is interesting that the standing-wave ratio and the dielectric constant are related linearly. This simple relationship does not hold between the reflection coefficient and the dielectric constant, however, hence the reflection coefficient is not considered further.

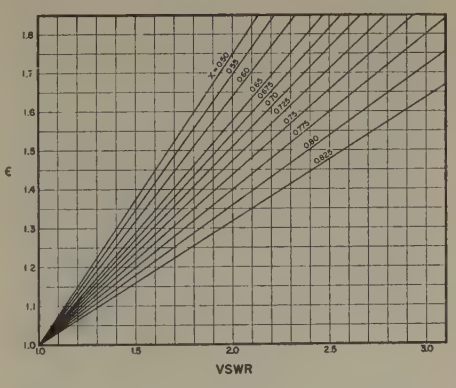

Fig. 1

In Fig. 1, the dielectric constant is plotted as a function of standing-wave ratio for various values of $x = \lambda/\lambda_c = f_c/f$. To use the curves, first calculate x from the operating and cut-off wavelengths or frequencies. This singles out one of the straight lines on the graph. Using the vswr it is desired to match out as the abscissa, read off the required value of ϵ on the ordinate scale. The slug is then made of material having this dielectric constant and should have an axial length

$$= \frac{\lambda}{4\sqrt{\epsilon - \left(\frac{\lambda}{\lambda_c}\right)^2}}.$$

As previously indicated, the match is then achieved by sliding the slug along the guide. If desired, the slug can be fixed in its final position by a transverse pin perpendicular to the E-vector or by other means.

For TEM systems, $\lambda_c = \infty$, and a simple result is obtained:

$$\epsilon = s$$

$$l = \frac{\lambda}{4\sqrt{\epsilon}}.$$

Additional curves may easily be added to the figure in the general case, or the range may be extended, by simply calculating a single point and joining it to the point 1, 1. Actually it is almost as simple to make the calculation directly from the formula in each specific case.

Present dielectric materials are available in steps of 0.05 from $\epsilon = 1$ to $\epsilon = 2.6$. Power handling capability is moderate with present materials due to accumulation of heat in the cellular structure of the dielectric. Contemplated improvements in these materials should reduce the loss tangent to a point where internal heating is almost negligible. For moderate standing-wave ratios this method should then be satisfactory even for very high power levels. For very large mismatches, and hence high dielectric constants, the method will not be suitable for high power levels even with perfect dielectrics, since breakdown between the dielectric and guide wall will become a factor. However, this might be overcome by pressurizing or by using two methods of matching in conjunction.

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Adventures in Calculus*

We have recently encountered two new mathematical results of interest:

$$\int \frac{dc}{\text{cabin}} = \log \text{cabin} + C = \text{house boat}$$

$$d(Hi\ Ho) = Hi\ dHo + Ho\ dHi.$$

M. S. Corrington
Radio Corp. of America
Camden, N. J.

* Received by the IRE, August 12, 1954.

* Received by the IRE, July 28, 1954.

Correspondence

Ratio Squarer*

During a recent symposium on military communications, a number of papers[1–4] discussed in part a new form of diversity reception identified as a "Ratio Squarer." Kindly reference and acknowledgment was made to the undersigned as the originator of this system, with a request for publication of the analysis which pointed the way to this development.

The following is accordingly a simplified analysis, which may be of interest to members of the Institute.

It was long felt that the sharper the diversity selection the better the operation of the system. Therefore, many workers have attempted to improve the sharpness of selection of diversity systems, and it is now possible to practically eliminate the weaker diversity signals and accept only the strongest with differences in diversity levels of less than a decibel. It is obvious that this mode of operation produces a signal-to-noise ratio equal to the best diversity signal at all times.

It appeared evident, however, to the writer that some thought should be given as to whether the goal of sharp diversity selection is a desirable one. The following analysis indicates that it is not desirable to completely discard the weaker diversity signals, because use may be made of them to improve the over-all signal-to-noise ratio. The maximum improvement, possible from this new expedient, is 3 db for two-receiver diversity, 4.8 db for three-receiver diversity and 6 db for four-receiver diversity. These figures are optimum values and occur only when the signals being received are equal in amplitude. When they differ from equality, the improvement decreases but at all times the combined signal-to-noise ratio is greater than the best of the individual diversity signal-to-noise ratio.

In order to determine the optimum diversity combining law, the following question must be answered. For a given ratio of diversity signal levels, how much of the weaker signal with its accompanying noise should be added to the stronger signal and its noise if an optimum signal-to-noise ratio is to be obtained? It is assumed that the signals are combined so that they add in phase, whereas the noise adds in a root-mean-square fashion. This may be readily accomplished by connecting the dc outputs of the signal demodulators in series.

Assumptions:

1. Signal from the diversity channels add linearly.
2. The noises add root-mean-square.
3. Two-receiver diversity is utilized.
4. The noise in one diversity receiver equals the noises from the other receiver.
5. The noise is random (white) in character.

D is the factor by which the weaker channels signal plus noise is divided in order to optimize the combined output signal-to-noise ratio. It is assumed that the gain of the stronger channel is unity and therefore D is equal to the gain of the stronger channel over the gain of the weaker channel.

To determine the value of D for maximum combined signal-to-noise ratio, it is necessary to differentiate the combined signal-to-noise ratio with respect to D. This differential is then set equal to zero in order to determine the optimum value of D.

$$S/N = \frac{S_1 + S_2/D}{\sqrt{N_1^2 + N_2^2/D^2}}. \tag{1}$$

But it is assumed that $N_2 = N_1$

$$S/N = \frac{S_1 + S_2/D}{\sqrt{N_1^2 + N_1^2/D^2}} \tag{2}$$

$$= \frac{DS_1 + S_2}{N_1\sqrt{D^2+1}} \tag{3}$$

$$\frac{dS/N}{dD} = \frac{N_1(D^2+1)^{1/2}S_1 - (DS_1+S_2)\frac{N_1}{2}(D^2+1)^{-1/2}2D}{N_1^2(D^2+1)} = 0 \tag{4}$$

$$(D^2+1)S_1 = (DS_1+S_2)D \tag{5}$$

$$D = \frac{S_1}{S_2}. \tag{6}$$

Where S_1 is the signal from one receiver and S_2 the signal from the second channel, N_1 is the noise accompanying S_1 and N_2 accompanying S_2.

It is therefore apparent that, in order to optimize the combined output signal-to-noise ratio in a system wherein the signal is made to add linearly and the noise root-mean-square, the ratio of the signal levels should be squared and then summated. In other words, if the signals were 4 db different in level, the ratio should be increased to 8 db before combination. Heretofore, it has been assumed that the weaker signal should be completely eliminated, but we see that this does not optimize the signal-to-noise ratio.

Fig. 1 shows a possible method for obtaining the ratio squarer selection curve (used by the U. S. Army Signal Corps in the AN/FGC-29 equipment). It is shown for a two-receiver diversity system, however, the basic arrangement may be readily adapted for three- or four-receiver diversity. It is seen that the signal and noise from receiver 1 are fed to a variable gain amplifier 1 which in turn feeds diode rectifier $D1$. The signal and noise from receiver 2 are fed to variable gain amplifier 2, which in turn feeds diode $D2$. Diode $D1$ and $D2$ are connected to a common load and the diode fed by the stronger signal conducts and biases the other diode off. Therefore, the output voltage across resistor R_{CL} is determined solely by the level of the stronger channel. This voltage is then fed through a lowpass filter and it is used to control the gain of the variable gain amplifiers 1 and 2. Thus, the ouput from the variable gain amplifier fed by the stronger signal is maintained constant. The gains of variable gain amplifier 1 and 2 are equal and there is no selection in this part of the circuit. It will be recognized that this is identical with the conventional common AVC diversity system.

The outputs of the individual variable gain amplifiers are fed to separate circuits which vary their gain linearly with signal

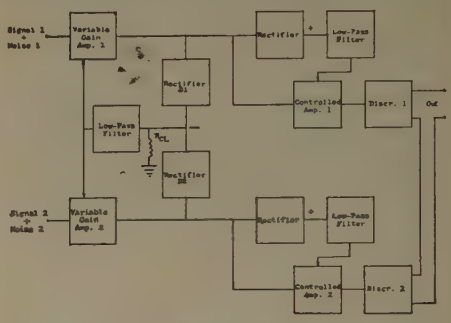

Fig. 1

level. By this process, the level difference between the weaker and stronger signals is squared. The manner in which this squaring is done is readily seen if one considers the circuit to be equivalent to a linear volume expander. A portion of the signal is rectified and the rectifier connected so that the bias on the controlled amplifier decreases with an increase in signal level. With this circuit, the gain may be made a direct function of the input level and therefore the ratio of the output of the two channels is proportional to the square of the input signal levels. The output of these amplifiers is then fed to demodulators (discriminators in the case of FSK reception) and then the outputs of the discriminators are connected in series giving the desired linear signal addition and root-mean-square noise addition effect.

Many tests were performed evaluating and comparing this system and the results were extremely gratifying. When compared with sharp selection-type diversity system and the common limiter diversity system, as well as other diversity systems, an appreciable reduction in error count was achieved and, in fact, a reduction in error count of 26 per cent was demonstrated over the next best system (at an error rate of approximately 2 errors per thousand characters).

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* Received by the IRE, July 13, 1954. This work was done as part of Contract No. DA36-039 sc-15359, administered by the Long-Range Section of Coles Signal Lab.

[1] A. Mack and R. H. Levine, "A New Multi-Channel Teletype Terminal for Use on Long Range HF Radio Systems," Signal Corps Eng. Labs., Ft. Monmouth, N. J.

[2] J. E. Boughtwood, "Circuit Design Aspects of Telegraph Terminal AN/FGC-29," Western Union Tel. Co., New York, N. Y.

[3] M. G. Crosby, "Optimum Diversity Combination," Crosby Labs., Inc., Hicksville, N. Y.

[4] F. H. Cusack, "Telegraph Terminal AN/FGC-29 Description and Application," Western Union Tel. Co., New York, N. Y.

Contributors

A. R. Bergen was born on April 7, 1923 in New York, N. Y. He received the degree of B.E.E. at the College at the City of New York in 1944, and the degree of M.S. at Columbia University in 1953. From 1944 to 1946 he served in the U. S. Signal Corps and later joined the engineering staff of the Westinghouse Electric Corporation. Since 1953, he has been in the Systems Analysis Group of the Electronics Research Laboratories at Columbia University.

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Mr. Bergen is a member of Sigma Xi, Tau Beta Pi, Eta Kappa Nu, and the AIEE.

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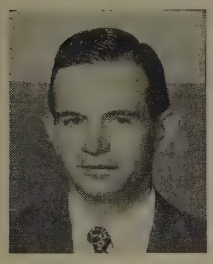
C. K. Birdsall (S'47–A'51) was born in New York, N. Y., on November 19, 1925. He received the B.S.E. and M.S.E. degrees in e.e. from the University of Michigan in 1946 and 1948 respectively, and the Ph.D. degree from Stanford University in 1951. At Stanford, Dr. Birdsall was a Sylvania Fellow in 1949, and an RCA Fellow in Electronics from 1949 to 1951.



C. K. Birdsall

Presently, he is a member of the technical staff of Hughes Research and Development Laboratories, Culver City, Calif., engaged in microwave tube research in the Electron Tube Laboratory. He is also Lecturer in Engineering, University of California, Los Angeles.

He is a member of Sigma Xi and Tau Beta Pi.

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H. J. Carlin (SM'50) was born in New York, N. Y., in 1917. He attended Columbia University, receiving the B.S. degree in 1938, and the M.S. degree in 1940. In 1947 he was awarded the D.E.E. degree from Polytechnic Institute of Brooklyn.

H. J. Carlin

Mr. Carlin was with Westinghouse from 1940 to 1945. In the latter part of 1945 he joined the Polytechnic Institute of Brooklyn, where he now holds the position of Research Professor. His major work has been in the fields of microwave devices and network theory.

He is a member of the AIEE, AAAS, Tau Beta Pi, Sigma Xi, and Eta Kappa Nu, and is Vice-Chairman of the IRE Professional Group on Circuit Theory.

E. A. Guillemin (A'41–SM'48–F'49), well-known authority on network analysis and synthesis, was born in Milwaukee, Wis., on May 8, 1898. He received the B.S. degree in e.e. in 1922 from the University of Wisconsin, and the S.M. degree in e.e. in 1924 from M.I.T. In 1926 he received the doctorate degree at the University of Munich on the Saltonstall Traveling Fellowship. He returned to M.I.T. as an instructor, becoming Assistant Professor in 1928, Associate Professor in 1936, and Prof. of Electrical Communications in 1944.

E. A. Guillemin

In 1940, he was appointed Consultant to the Microwave Committee of the NDRC. In this capacity, approximately half of his time was devoted to consultation with various groups in the Radiation Laboratory, M.I.T., on a variety of problems dealing with the design of electrical networks for special applications. Of significance was the development of a network for the production of radar pulses. In 1941, Dr. Guillemin took over the administrative responsibilities of the Communications Option, Dept. of Electrical Engineering at M.I.T.

In 1948, Dr. Guillemin was awarded the President's Certificate of Merit for his outstanding wartime contributions. He is particularly known for two volumes, "Communications Networks," and a reference entitled, "The Mathematics of Circuit Analysis." He is a Fellow of the AIEE and a member of the ASEE.

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F. Haber (S'48–A'49) was born in New York, N. Y., on July 1, 1921. He attended City College of New York and Pennsylvania State University, receiving the B.S.E.E. degree from the latter in 1948. His undergraduate studies were interrupted by military service in the U. S. Army. In 1953 he received the M.S.E.E. degree from the University of Pennsylvania.

F. Haber

Mr. Haber joined the Moore School of Electrical Engineering in 1951, where he now holds the rank of Associate. At present he is engaged in research on interference in communications systems, and part-time teaching. During 1954 he was a nonresident instructor at M.I.T., conducting a course in electrical communication. Previously, from 1948–1951, he had been with RCA Victor, Camden, N. J., and in 1948 with the Arma Corporation, Brooklyn, N. Y.

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W. P. Mason (A'35–F'43) was born in Colorado Springs, Colo., in 1900. He graduated from the University of Kansas in 1921, and obtained the A.M. and Ph.D. degrees from Columbia in 1924 and 1928 respectively. He is the author of two books, "Electromechanical Transducers and Wave Filters," and "Piezoelectric Crystals and Their Applications to Ultrasonics."

W. P. Mason

Since 1921 to the present, Dr. Mason has been a research engineer and physicist with the Bell Telephone Labs., where he is now head of the mechanics research group. His principal contributions have been made in the fields of piezoelectric crystals, ultrasonics, and mechanics.

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M. W. Muller was born in 1922 in Vienna, Austria. He received the B.S. degree in e.e. in 1949, and the A.M. degree in physics in 1952, both from Columbia University.

M. W. Muller

Mr. Muller served in the armed forces from 1943 to 1946. He worked as a research associate at Columbia Radiation Lab. until 1952, engaged in research connected with magnetrons. Since 1952 he has been with Varian Associates, where he is concerned with problems of electron dynamics and microwave tubes.

Mr. Muller is a member of the APS, Tau Beta Pi, and Sigma Xi.

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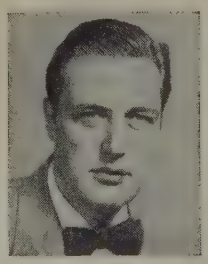
B. M. Oliver (S'40–A'40–M'46–SM'53–F'54) was born May 27, 1916, at Santa Cruz, Calif. He received the B.A. degree in e.e. from Stanford University in 1935, and the M.S. degree from the California Institute of Technology in 1936. Following a year of study in Germany under an exchange scholarship, he returned to the Institute, receiving the Ph.D. degree in 1940.



B. M. Oliver

From 1939 to 1952 he was employed at the Bell Telephone Labs., in television research and radar development. He is now director of research at the Hewlett-Packard Company.

Contributors

Robert A. Pucel (S'48–A'52–S'54) was born in Ely, Minnesota in 1926. After serving in the U. S. Navy from 1945 to 1946, he entered the Massachusetts Institute of Technology and received the B.S. and M.S. degrees in electrical engineering, in 1951. Under the cooperative plan in electrical engineering, during the years from 1948 to 1951 he worked at the General Electric Company.

R. A. Pucel

Upon graduation, Mr. Pucel joined the Research Division of the Raytheon Manufacturing Company, Waltham, Massachusetts, where he was a member of the microwave noise study group. In 1952, he returned to M.I.T. as a doctoral candidate. At present, he is a staff member of the Research Laboratory of Electronics at M.I.T., and works part-time at the Research Division of the Raytheon Manufacturing Company on the study of transistors.

Mr. Pucel is an associate member of Sigma Xi.

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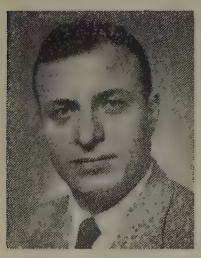
J. R. Ragazzini (A'41–M'46–SM'52) was born on January 3, 1912 in New York City, N. Y. He received the degrees of B.S. and E.E. at the College of the City of New York in 1932 and 1933, and the A.M. and Ph.D. degrees at Columbia University in 1939 and 1941. After service with one of the New York City Departments, he became an instructor in the School of Technology of the City College. During World War II he was a project director in the Division of War Research at Columbia University, and has been a consultant in the fields of servomechanisms, analog computers, and noise problems. He was Director of the Electronics Research Laboratories at Columbia until 1953, and is now a member of its research staff. Also, he joined the faculty of Columbia University in 1941, where he is now Professor of Electrical Engineering and Executive Officer.



J. R. Ragazzini

He is a member of Phi Beta Kappa, Sigma Xi, Tau Beta Pi, Eta Kappa Nu, the AIEE, and the ASEE.

❖

Hermann Statz was born in Germany on January 9, 1928. He received the Diploma in Physics in 1949, and the Dokter der Naturwissenschaften degree in 1951, from the Technische Hochschule, Stuttgart. During the years from 1949 to 1951 he was a research associate at the Max Planck Institut für Metallforschung, Stuttgart. From 1951 to 1952 he held a research stipend from Deutsche Forschungsgemeinschaft; his work was concerned with problems in theoretical solid state physics. In 1952 he joined the staff of the Solid State and Molecular Theory Group at the Massachusetts Institute of Technology, where he worked on the theory of ferromagnetism. Since 1953 he has been a staff member of the Research Division of the Raytheon Manufacturing Company, Waltham, Massachusetts, where he is doing research in solid state physics and related fields.

H. Statz

Dr. Statz is a member of the American Physical Society.

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Leo Storch was born on March 3, 1921, in Vienna, Austria. In January, 1944, he received the B.E.E. degree, cum laude, from the School of Technology, College of the City of New York. He was awarded the M.S. degree by the Graduate School of Stevens Institute of Technology in June, 1947.

Leo Storch

From 1944 to 1947, Mr. Storch worked as a development engineer with the Western Electric Company, Kearny, N. J. He was engaged in the circuit design for a wide variety of testing and measurement equipment. Subsequently, he transferred to RCA Victor, Camden, N. J., as advanced development engineer in the aviation radio department, where he was primarily concerned with the development of antenna impedance-matching networks and automatic tuning systems for wide-band aircraft communications transmitters.

In 1950, Mr. Storch became a member of the Research and Development Laboratories of the Hughes Aircraft Company, Culver City, California, where he is active in the development of specialized communications receivers and networks.

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J. W. Sullivan was born on March 15, 1926, in Mansfield, Ohio. In 1952 he received the B.E.E. and the M.S. degrees from Ohio State University.

During the Second World War Mr. Sullivan served as a radar technician in the United States Marine Corps. Since 1952 he has been employed as a member of the technical staff of the Bell Telephone Laboratories, at Murray Hill, New Jersey where his work concerns vacuum tube research.

J. W. Sullivan

In addition to his other activities Mr. Sullivan is a member of Sigma Xi, and Tau Beta Pi.

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D. E. Thomas (A'47–SM'47) was born on May 12, 1907, in Hanover, Pa. He received the B.S. degree in e.e. from Pennsylvania State University in 1929, and the M.A. degree from Columbia University in 1932.

D. E. Thomas

He joined Bell Telephone Labs., New York, in 1929, where he first was engaged in work on submarine cable, and then on seaborne and airborne radar equipment. In 1942 he entered the Signal Corps, and was assigned to the staff of the Chief Signal Officer in charge of Radio Countermeasures. In 1946, on his return to Bell Telephone Labs., as a member of the technical staff, he became active in the development and installation of the first deep-sea repeatered submarine telephone cables which went into service between Key West and Havana in 1950. Since then he has devoted his time to the development of transistor devices and circuits. Mr. Thomas was a civilian member of the Panel on Electronic Countermeasures of the Res. and Dev. Board, Dept. of Defense, from 1948 to 1953.

He is a member of Phi Kappa Phi and Tau Beta Pi.

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R. F. Wick was born in Creston, Ia., in 1907. He received the B.S. degree from the A. & M. College of Texas in 1928, and taught part-time in the electrical engineering department for two years, receiving the M.S. degree in 1930.

R. F. Wick

Since 1930 Mr. Wick has been with the research department of the Bell Telephone Labs., engaged in special electrical networks design, and, more recently, in ultrasonics and ferroelectric research.

He is a member of the Acoustical Society.

IRE News and Radio Notes

Chicago Section Presents IRE Awards at Recognition Banquet

Climaxing its 1953–54 year the Chicago Section held a Spring Recognition Banquet, honoring IRE award winners in the Chicago area, thus launching with conspicuous success the new procedure whereby the presentation of Fellow Awards is made by the recipient's Section rather than at the IRE Convention.

The banquet paid tribute to William L. Everitt, past president of IRE and winner of the 1954 IRE medal of honor. Other honored guests received Fellowship awards, among them William L. Dunn, for his contribution in the development of radio receivers; Kurt Schlesinger, for work on cathode-ray tubes and television circuitry; Wilber L. Webb, for early contributions and engineering leadership in radio direction finding, radar, and radio communication.

Four students from the Chicago area received the 1954 Student Awards. They were: Philip Lopresti, Notre Dame; John Miyasaki, Illinois Institute of Technology; John Larney, Northwestern; Robert Hardin, Illinois.

Guests of honor at the event were: William Hewlett, president of IRE; George Bailey, executive secretary; Laurence G. Cumming, technical secretary; William Copp, advertising manager of the PROCEEDINGS; John Ryder, director-at-large; Charles Marshall, director of Region 5; Ernst Weber and Alois Graf, vice chairmen of the National PG Committee.

(*Above, back row*) Guests of honor Alo Graf, Charles Marshall, William Hewlet (*Front row*) Laurence Cumming, Albe Gerlach, William Everitt, George Bailey

President W. R. Hewlett presents Fellow Certificates to Kurt Schlesinger and William Dunn

(*Below*) J. D. Ryder presents Student Awards to J. M. Larney, J. K. Miyasaki, P. V. Lopresti, and R. H. Hardin. (*Back row*) Looking on are A. H. Wing, Jr., G. T. Flesher, H. Ellithorn, P. H. Schwarzlose, IRE Representatives, and C. J. Marshall, Director of Region 5.

IRE News and Radio Notes

IRE Board Announces 1955 Award Winners

Harold T. Friis, Director of Radio Research, Bell Telephone Laboratories, will receive the 1955 IRE Medal of Honor, the Institute's highest award. The award will be given to Mr. Friis "for his outstanding technical contributions in the expansion of the useful spectrum of radio frequencies, and for the inspiration and leadership he has given to young engineers."

The Morris Liebmann Memorial Prize, awarded annually to an IRE member recently making an important contribution in the field, will be given to Arthur V. Loughren, Director of Research, Hazeltine Corp., "for his leadership and technical contributions in the formulation of the signal specification for compatible color television."

Bernard Salzberg, Naval Research Laboratory, will receive the Harry Diamond Memorial Award, given persons in government service for outstanding work in radio or electronics. The award will be presented "for his contributions in the fields of electron tubes, circuits, and military electronics."

The Vladimir K. Zworykin Television Prize will go to Harold B. Law, RCA Laboratories, "for development of techniques and processes resulting in a practical form of shadow-mask tri-color kinescope."

The awards will be presented at the Annual IRE Banquet during the 1955 National Convention next March.

ASTM Standards on Tube Materials Available

The American Society for Testing Materials has recently issued standards on "Materials for Radio Tubes and Electronic Devices." The standards cover specifications, tests, and recommended practices and are available from the ASTM, 1916 Race St., Philadelphia, Pa., at $2.75 per copy.

Participants at Broadcast Symposium. From left to right: G. F. Leydorf, session moderator; L. Katz and C. B. Plummer, speakers; L. Winner, symposium chairman; J. Epstein and L. E. Rawls; speakers; J. R. Wentworth, recipient of Resnik Foundation plaque for outstanding contributions to ultrahigh broadcasting.

PG on Broadcast Transmission Systems Holds Symposium

The Professional Group on Broadcast Transmission Systems held their fourth annual symposium, September 24 and 25, at the Carter Hotel in Cleveland, Ohio. The moderators were Carl Smith, Joseph Epperson and George Leydorf, and the talks covered many phases of Broadcast Transmission. The speakers were: Rodney Chipp, W. P. Bennett, L. L. Koros, Frank Bias, Charles Page, M. G. Scheraga, Robert Dickert, Graydon Lloyd, John Barstow, Curtis Plummer, Louis Katz, Jess Epstein, J. R. Whitworth, L. E. Rawls.

On Friday evening, cocktails preceded the award banquet, at which George Brown was guest speaker. John R. Whitworth received a plaque from the Resnick Foundation for his contributions to ultrahigh broadcasting. The symposium closed the following day with additional talks and a final "information please" roundtable.

Calender of Coming Events

IRE Baltimore Section-PGANE East Coast Conference on Airborne and Navigational Electronics, Sheraton-Belvedere Hotel, Baltimore, Md., November 4–5

IRE-PIB Microwave Symposium, Engineering Societies Auditorium, New York, N. Y., November 8–10

IRE-AIEE Conference on Electrical Techniques in Medicine and Biology, Morrison Hotel, Chicago, Ill., November 10–11

IRE Quality Control Symposium, Statler Hotel, New York, N. Y., November 12–13

Symposium on Fluctuation Phenomena in Microwave Sources, Western Union Auditorium, New York, N. Y., November 18–19

IRE Kansas City Section Annual Electronics Conference, Hotel President, Kansas City, Mo., November 18–19

IRE-AIEE-ACM Eastern Computer Conference, Bellevue-Stratford Hotel, Philadelphia, Pa., December 8–10

Society for Industrial and Applied Mathematics, first national meeting, University of Pittsburgh, December 27–29

IRE-AIEE-NBS-URSI Conference on High Frequency Measurements, Hotel Statler, Washington, D. C., January 17–19

RETMA Symposium on Printed Wiring and Circuitry, University of Pennsylvania, Philadelphia, Pa., January 20–21

IRE-IAS-ION-RTCA Electronics in Aviation Symposium, Hotel Astor, New York, N. Y., January 26

AIEE Winter General Meeting, Hotel Statler, New York, N. Y., January 31–February 4

1955 Southwestern IRE Conference and Electronics Show, Baker Hotel, Dallas, Tex., February 10–12

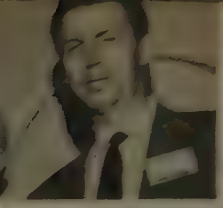
IRE National Convention, Waldorf-Astoria Hotel and Kingsbridge Armory, New York, N. Y., March 21–24

IRE-PIB Symposium on Modern Network Synthesis, Engineering Societies Building, New York, N. Y., April 12–15

SMPTE 57th Semiannual Convention, Hotel Drake, Chicago, Ill., April 17–22

Semiconductor Symposium, Electrochemical Society, Cincinnati, Ohio, May 2–5.

1954 WESCON Entertains Record Crowd



(*Above*) WESCON and IRE VIP'S look over the show. L to r: George W. Bailey, executive secretary, IRE; W. D. Hershberger WESCON chairman; Laurence G. Cumming, technical secretary, IRE; Thomas P. Walker, WESCON vice-chairman; and Mal Mobley, Jr., WESCON business manager.

More than 23,000 engineers, scientists, technicians, and members of allied industries registered this August 25–27 at Los Angeles for the 1954 WESCON (Western Electronic Show and Convention). Over 2,800 people attended the technical sessions —held in the Ambassador Hotel and the Institute of Aeronautical Sciences Building —where 115 papers were delivered. Co-sponsors of the event were the Los Angeles and San Francisco Sections of the IRE and WCEMA (West Coast Electronic Manufacturers' Association).

Highlight of WESCON was the All-Industry Luncheon, presided over by Dr. W. D. Hershberger, chairman, and featuring William R. Hewlett, national IRE president, who spoke on "Why Electronics Grows with the West." Dr. John B. Smyth, of the U. S. Navy Electronics Laboratory in San Diego, received the IRE Seventh Region Annual Achievement Award. The annual WCEMA scholarship awards were made to outstanding students of eleven West coast colleges and universities. Paul H. Tartak, chairman of the scholarship committee presented the awards to:

Richard W. Martin and Raymond H. Strong, California State Polytechnic College; Elbert Hartsfield and Jack W. Ward, UCLA; Angel Arellano, University of California; Ronald A. Schrock, Oregon State College; Raymond L. Strauss, Edward A. Starr, and K. H. Wilson, San Diego State College; Daniel S. Boessow, University of Santa Clara; Frank Glass and W. D. Hightower, San Jose State College; B. T. Preston, Stanford University, J. T. Anreika, University of Washington; and Robert E. England, USC. The recipient of the award made by the California Institute of Technology will be announced later.

(*Above*) John [illegible] winner of the IRE Seventh Region Achievement Award.

The technical program included sessions on telemetering, component parts, antennas, analog devices and computing systems, computer circuits and components, circuit design, microwave antennas, microwave theory and techniques, engineering management, vehicular communications, antennas and propagation, electron tubes, broadcast and television receivers, systems analysis, information theory, airborne electronics, audio, electron devices, and ultrasonics. In addition, a panel discussion, "Problems in Vehicular Communications," was moderated by F.C.C. Commissioner E. M. Webster.

Vice-chairmen for WESCON were C. Frederick Wolcott and Thomas P. Walker. Other members of the 1954 board of directors were Secretary-Treasurer, Leon B. Ungar, Noel E. Porter, Richard A. Huggins, Joseph H. Landells, and Walter E. Noller.

Committee chairmen and alternates were: *Technical Program*, Frederick G. Suffield, V. J. Braun; *Arrangements*, John K. Gossland, Glyn A. Neff; *Special Events*, Richard F. Walz; *Women's Activities*, Mrs. James N. Whitaker, Mrs. Elmer P. Gerwick; *All-Industry Cocktail Party*, W. Bert Knight, Richard G. Leitner; *All-Industry Luncheon*, Walter E. Peterson, William A. Bowen, Jr.; *Field trips*, David T. Wersen, William S. Klein; *Show Registration*, R. V. Weatherford, Thomas A. Lynch; *Convention Registration*, Burgess Dempster & G. La Casse; *Visitors Service*, Norman Neely, George Davis; *Visitors Housing*, Granger Yarbrough; Ernst H. Schreiber, *Publicity*, Don Larson, and Edward Bernsley.

Chairmen of the Los Angeles and San Francisco Sections of IRE were Bruce S. Angwin and A. J. Norris respectively; the Regional Director was Dr. Joseph Perrin. WCEMA was headed this year by E. P. Gerwick and J. J. Halloran.

Pan Pacific Auditorium where more than 600 manufacturers of electronic equipment exhibited in 522 booths. A two-way TV intercom, silicon transistors, color television receivers, ultrasonic cleaners, high fidelity equipment and computers were among the products demonstrated.

Eastern Joint Computer Conference to Meet Next Month

"Design & Application of Small Digital Computers" is the theme of the Eastern Joint Computer Conference and Exhibition to be held at the Bellevue-Stratford Hotel, Philadelphia, Decmber 8–10, 1954.

The program includes papers on computer comparisons, input-output devices, computer systems and characteristics, mathematics, programming, business and scientific applications. Three panel discussions on business applications, redundancy checking, and small computers as large computer auxiliaries are scheduled for Thursday afternoon. Inspection trips will be conducted to the Bell System automatic message accounting center, Burroughs Research Center, the Franklin Institute network calculator, and Remington Rand's Eckert-Mauchly Division. Exhibits of at least 60 companies will occupy the entire 18th floor of the hotel. A cocktail party will be held Wednesday evening.

Preliminary programs and registration material have been mailed to all previous registrants, ACM members, and many AIEE & IRE members. Others may register at the conference, or request material by writing P.O. Box 7825, Phila. 1, Pa.

K. A. Norton Presented Stuart Ballantine Medal

A Stuart Ballantine Medal, awarded by The Franklin Institute of the State of Pennsylvania, for outstanding achievement in the field of communication was presented to Kenneth Alva Norton, Chief, Radio Propagation Engineering Division, National Bureau of Standards. The award was made at the Institute's annual Medal Day ceremonies, Wednesday evening, October 20. Dr. Henry B. Allen, executive vice president and director, said Mr. Norton is being honored "in consideration of his contributions over a period of twenty-five years in the field of radio propagation through which our knowledge has been considerably increased by his measurements; our insight broadened by his theoretical work; our engineering calculations made easier by his charts and our broadcasting frequency allocations above 50 mc established more effectively through his guidance."

Born in Rockwell City, Iowa, February 27, 1907, Mr. Norton received the B.S. degree from the University of Chicago in 1928. In 1929 he was employed as an inspector in Western Electric's Chicago office. From 1929–1934 he was a junior physicist, radio section, Bureau of Standards. For an eight-year period, 1934–1942, he served as radio engineer for the Federal Communication Commission, Washington, D. C. From 1942–1943 he was assistant director of the operational research group, Office of Chief Signal Officer. He was affiliated from 1943–1944 with the operational research section, Eighth Air Force, England in the capacity of radio and tactical countermeasures analyst. Subsequently, until 1946, he was a radio propagation consultant, Office of Chief Signal Officer, Washington, D. C.

Mr. Norton is a fellow of the Physical Society, a member of the Institute of Electrical Engineers, the Mathematical Society, and the Statistical Association.

Professional Group News

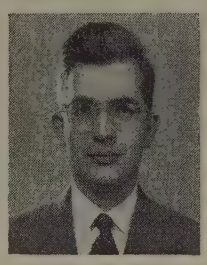
New Chapters Approved

The IRE Executive Committee at its meeting on September 8th, 1954, approved the petitions for the establishment of the following Chapters. San Francisco Chapter—Professional Group on Engineering Management, Warren W. Chan, 395 Page Mill Rd., Palo Alto, California, organizer. Cleveland Chapter—Professional Group on Industrial Electronics, Carl E. Smith, 4900 Euclid Ave., Cleveland, Ohio, organizer. Urbana, Illinois (Chicago Section) Chapter—Professional Group on Circuit Theory, Ray M. Wainwright, School of E. E., University of Illinois, Urbana, Ill., organizer.

Obituaries

Armin J. Herz (M'49–SM'53), patent attorney in Chicago, Illinois, died recently. From 1925 to 1938, Mr. Herz was with the firm of Gustav Ganz and Co. as a patent lawyer. He received the M.E. degree from Vienna Technical University in 1921.

William G. Tuller (S'37–A'40–M'45–SM'48), Vice President in charge of Engineering, and a Direc**or** of Melpar, Inc. (a subsidiary of Westinghouse Air Brake Company), one of the 28 victims of the KLM plane crash at Shannon, Ireland on September 5, was born in Rutherford, New Jersey, September 8, 1918. He was a pioneer in the field of radar and was one of the six scientists who developed microwave radar. At the end of the war, nearly every ship in the United States Navy carried at least one radar whose components were developed by Dr. Tuller or the group under his direction. For his efforts during the war, he was awarded the United States Navy's Certificate of Commendation.



W. G. Tuller

During the postwar period Dr. Tuller was engaged in research on the basic theory underlying the transmission of information over a wire or radio circuit. He developed an equation setting forth the limit on the rate at which information could be sent over such a circuit, and defining a practical measure of information. This forms the basis of the "information theory" that is, in many aspects, revolutionizing the electronics industry.

Prior to joining Melpar, he worked for the Equipment Division of Raytheon and later with the M.I.T. Research Laboratory of Electronics. He became associated with Melpar, Inc. in 1947 as a Project Engineer and rose rapidly to Chief Engineer, Director of Engineering, and finally Vice President in Charge of Engineering.

He obtained his doctor's degree in E.E. from M.I.T. in 1948, was elected to Sigma Xi, actively participated in the affairs of the IRE, was a member of several of its committees, and Chairman of the Professional Group on Information Theory.

Technical Committee Notes

Under the chairmanship of Henry Burkhard the **Facsimile** Committee convened on July 9th. In connection with a search for a method to reproduce the facsimile test chart, B. H. Klyce showed a large colored print made by the collotype printing process. Mr. Klyce will obtain additional information on the process and also will check approximate cost of reproducing the chart by this means. L. R. Lankes sent K. R. McConnell a letter suggesting possible modification of the test chart and information on photographic means for its reproduction. In an answering letter, Mr. McConnell commented on the components of the chart and forwarded information on the Munsell density steps.

The **Feedback Control Systems** Committee convened on August 5th under the Chairmanship of J. E. Ward. The Committee discussed the definition of "loop return signal" and "loop difference signal." Chairman Ward reported on the formation of the Professional Group on Automatic Control Technology. Its petition is expected to be approved by the Executive Committee in the fall. Chairman Ward distributed copies of proposed definitions prepared by S. B. Williams and G. A. Biernson. The definition of terms will become the Subcommittee's work, and definitions proposed by Committee members, as a result of assignments, will be turned over to the Subcommittee. The Committee will consider questions on Measurements and Standards of Performance. Terms resulting from such discussions will be turned over to the subcommittee also. Many redundant terms now in use may be thus eliminated.

On August 31st the **Recording and Reproducing** Committee met under the chairmanship of A. W. Friend and the membership of each Subcommittee was read to those present. R. C. Moyer was introduced as the new Chairman of Subcommittee 19.1. Dr. S. J. Begun offered some suggestions to Mr. Moyer in regard to his subcommittee's work. E. W. D'Arcy and F. A. Comerci asked to become members of Subcommittee 19.1. A. Zuckerman of RETMA was introduced to the committee and presented his viewpoint on matters relating to standardization. The entries proposed for the Master Index of Measurement Standards were discussed. Mr. H. Davies of the British Broadcasting Corporation was introduced to the committee, and the meeting was left open for discussion with Mr. Davies.

The **Symbols** Committee met on August 26th under the chairmanship of K. E. Anspach, who welcomed R. W. Stern as a new member. The committee noted with regret that A. G. Clavier had resigned. Dr. Harry Stockman, who had sent in comments on the recently issued Standards on Graphical Symbols, discussed that topic with the committee. M. P. Robinson was appointed Chairman of Subcommittee 21.2 to replace C. D. Mitchell. The excellent work of Mr. Mitchell, during the difficult period when many varieties of symbols were being introduced, was noted and an expression of thanks unanimously voted by the Committee. A. C. Reynolds, Jr., made a report on the activity of Subcommittee 21.3 and F. M. Bailey reported on Subcommittee 21.4.

Abstracts of Transactions of the I.R.E.

The following issues of "Transactions" have just been published and are now available from The Institute of Radio Engineers, Inc., 1 East 79 Street, New York 21, N. Y. at the following prices. The contents of each issue and, where available, abstracts of technical papers are given below.

Sponsoring Group	Publication	Group Members	IRE Members	Non-Members*
Audio	Vol. AU-2, No. 5	$ .70	$1.05	$ 2.10
Circuit Theory	Vol. CT-1, No. 3	1.70	2.55	5.10
Electronic Computers	Vol. EC-3, No. 3	1.80	2.70	5.40
Information Theory	PGIT-4	3.35	5.00	10.00
Radio Telemetry and Remote Control	PGRTRC-1	.85	1.25	2.55

*** Public libraries and colleges can purchase copies at IRE Member rates.**

AUDIO

Vol. AU-2, No. 5, September-October, 1954

PGA News

Magnetic Recorders for Data Recording Under Adverse Environments—G. L. Davies

Describes types of magnetic recording equipment specifically developed for multichannel recording of data information such as vibration, displacement, pressure, temperature, etc. Equipment has been designed to operate over wide ranges of temperatures, pressures, humidities, accelerations, and vibrations as experienced in aircraft, vehicles, and other adverse environments.

Magnetic Recording—A Report on the State of the Art—W. T. Selsted and R. H. Snyder

This paper describes many of the present recording techniques and the limitations of each. Limitations in regard to performance such as frequency response, signal-to-noise ratio, and speed stability are considered from the standpoint of the machine as well as the medium used in the recording process. Many of the present day applications of magnetic recording are discussed, including those of television recording and the application of magnetic recording to seismic exploration.

Cathode Bias Resistor for Class A Triode—M. R. Winkler

This paper presents a useful "handbook type" equation for calculating a suitable self-bias resistor for a Class A triode amplifier stage designed for maximum signal output. It is shown also that the loss in gain is six decibels, when the cathode resistor is not shunted by a by-pass capacitor.

This information provides a timesaving method for determining the cathode resistor, with only a knowledge of plate resistance and transconductance, and without the necessity of a family of plate characteristic curves.

Navy Standardization of 1/4 Inch Magnetic Tape and Recorder-Reproducers—F. A. Comerci, S. Wilpon, and R. Schwartz

Standardization of magnetic tape for use by Federal agencies has been facilitated through the preparation and distribution of a "standard reference tape." Similar standardization of recorder-reproducers has been made possible through the use of a "standard tape recording." The properties of the tape, the material recorded, and the purpose and use of the tapes are given in detail.

CIRCUIT THEORY

Vol. CT-1, No. 3, September, 1954

The International Scientific Radio Union—J. M. Pettit (Editorial)

The Approximation Problem—W. H. Kautz (Foreword)

The Approximation Problem of Network Synthesis—S. Winkler

The approximation problem of network synthesis is surveyed with extensive reference to the recent literature, encompassing both the frequency and the time domains. The topics reviewed include the concept of complex frequency, approximation by point-coincident polynomials, and the Fourier series method. Butterworth functions for maximally-flat approximation and Tchebycheff polynomials for equal-ripple approximation inside the pass band are treated, as well as Jacobian elliptic functions which provide Tchebycheff behavior both inside and outside the pass band. A detailed review of the literature of the potential analog is given with reference to applications of the analogy and to the use of electrolytic tanks and rubber sheets. Mention is made of continued fractions and Laguerre polynomials together with references to papers discussing the synthesis of networks with prescribed transient response. A bibliography of 240 items is included.

A General Method for Time Domain Network Synthesis—F. B. Hli

The method presented in this paper solves the general problem of synthesis in the time domain, that is, given an arbitrary time function as the specified input to an electrical network, and another such time function as the desired output, it shows how to go about designing the network. These arbitrary time functions may be prescribed either graphically, or as analytic functions, or merely as sequences of values at stated intervals of time.

The first step is to write the input and output time functions in time-sequence form. Then synthetic division of the output time sequence by the input time sequence yield the sequence of areas under the impulse curve. Using the power series expansion of the Dirichlet series representation of the Laplace transform, it is possible to get the network system function from the above sequence of areas as a rational function.

This rational function gives a good approximation in the time domain and is easily computed. Examples are included to illustrate the method.

Transient Synthesis in the Time Domain—W. H. Kautz

One of the numerous available methods for synthesizing a network to have a specified transient response is described, with emphasis placed on the actual procedure of solution. This method, which is based upon the use of orthogonal exponential functions, is carried out principally by time-domain rather than the more common frequency-domain operations. It enjoys broader application than most other solutions to the transient synthesis problem, each of which suffers from one or more of several disadvantages: lack of control over the approximation error in time; severe mathematical complexity; networks with many more circuit elements than are necessary; or networks conveniently realizable only in a fixed form (e.g., lossless lattices, resistance-capacitance chains, networks without parasitic capacitance, etc.). Several examples are presented.

The Introduction of Constraints into Feedback System Designs—J. H. Wescott

Networks are frequently designed on the basis of minimizing a measure of error to a given specification. The physical limitations of practical circuit elements can often be introduced into such problems by means of constraints placed on the form of the solution. This paper presents a new way of introducing such constraints and reviews some examples where constraints have been successfully employed.

The constraints discussed may be divided into two groups; namely, those depending on some isolated feature of the frequency characteristic of the solution, such as the dc gain or the parasitic-capacitance high-frequency asymptote, and those depending on the complete frequency characteristic, such as the power-handling capacity of an output stage. It is suggested that the introduction of constraints may require a change in the basis of error assessment. In this connection, an example is given based on a time-weighted criterion.

A review of the statistical method of analysis of signals is included, with a critical discussion of the definition of the spectral density function, which forms the basis of the statistical methods of circuit design.

Numerical Analysis for Network Design—J. G. Truxal

The numerical-analysis techniques appropriate for the rapid approximate analysis of network response can be unified in terms of the z transform, or generalized Laplace transform. Application of the methods of analysis requires knowledge of only the z-domain integration operators for the multiple integration of impulse trains. The time functions can be approximated by rectangular pulses, straight lines, or parabolas of any order.

The numerical methods are illustrated with several examples. The transient response of a network can be evaluated without factoring a polynomial or adding damped sinusoids; the closed-loop transient response of a feedback system can be determined directly from the open-loop characteristics; a system with a transportation lag can be analyzed; the relations between frequency-domain and time-domain characteristics can be simplified; and a variety of other analysis problems are rapidly and easily considered. The same numerical methods can be utilized to determine the transfer function of a network for a prescribed transient response.

A Note on Filter Synthesis—B. J. Bennett

An equal-ripple type of approximation to constant attenuation in both stop and pass bands is often desirable in electric filter design. Elliptic functions have been extensively employed in solutions of problems of this sort. Double-entry tables are required and the mathematics is, in general, involved.

Transient Response of Cascaded Tuned Circuits—A. I. Talkin

The transient response of two cascaded stagger-tuned amplifier stages to a pulse of arbitrary carrier frequency is derived, starting from the response of a single-tuned circuit. The results are obtained by solving the differential equation of the network, and by employing a

simple stratagem which immediately yields the arbitrary constants.

The result is extended to three stagger-tuned stages, and the extension to an arbitrary number of stages is indicated. The transient response of two synchronously-tuned stages is also derived, based on the solution of the stagger-tuned case. No approximations are made in any of the derivations.

On the Bott-Duffin Synthesis of Driving-Point Impedances—V. Belevitch (Correspondence)

A Low-Pass Transformation for Z-Transforms—W. H. Huggins (Correspondence)

Error-Criterion vs. Harmonic Content—C. H. Page (Correspondence)

What is Nature's Error Criterion—E. A. Guillemin (Correspondence)

PCGT News

ELECTRONIC COMPUTERS

VOL. EC-3, NO. 3, SEPTEMBER, 1954

The Editor

A Permanent High Speed Store for Use With Digital Computers—R. D. Ryan

A new type of high speed store is proposed for an electronic digital computer using interpretive program techniques. The store is based on the flying spot technique used in television signal generation. The information in the store may be read rapidly but is nonerasable. This story has the advantages of high storage density, good reliability and nonvolatility of information.

Application of Boolean Algebra to Switching Circuit Design and to Error Detection—D. E. Muller

A solution is sought to the general problem of simplifying switching circuits that have more than one output. The mathematical treatment of the problem applies only to circuits that may be represented by "polynomials" in Boolean algebra. It is shown that certain parts of the multiple output problem for such circuits may be reduced to a single output problem whose inputs are equal in number to the sum of the numbers of inputs and outputs in the original problem. A particularly simple reduction may be effected in the case of two outputs.

Various techniques are described for simplifying Boolean expressions, called "+polynomials," in which the operation, "exclusive or," appears between terms. The methods described are particularly suitable for use with an automatic computer, and have been tested on the Illiac.

An unexpected metric relationship is shown to exist between the members of certain classes of "+polynomials" called "nets." This relaship may be used for constructing error-detecting codes, provided the number of bits in the code is a power of two.

An Algebraic Theory for Use in Digital Computer Design—E. C. Nelson

An algebraic theory of the logical operation of digital computers is developed. This theory takes into account the dynamic (time) behavior of computer processes. The computer signals and computer elements are described. Their properties which are pertinent to the logical operation of digital computers are abstracted and formulated in mathematical terms. The signals are represented by algebraic symbols, and the way they are transformed by the elements of the computer is represented in terms of algebraic operations and functions. This computer algebra is based on Boolean algebra. Time is treated as a discrete variable and a method of taking into account the time relationships in the computer processes is developed. Specific components, such as gates, flip-flops, and magnetic drums, are analyzed, and an algebraic description of their operation is obtained.

An Improved Reading System for Magnetically Recorded Digital Data—Samuel Lubkin

In magnetic recording of pulses, whether on drum or tape, the resulting flux pattern is affected by proximity of adjacent pulses. The best defined region is that adjacent to the maximum. In reading, the signal is the derivative of the flux. In the best defined region, this is close to a straight line passing through zero when the flux is maximum. The slope of the curve at the zero changes sign with change of pulse polarity. A new method of reading is described which examines the signal from the head for such transitions from positive to negative or reverse as indications that a positive or negative pulse had been recorded. This is done by gating the inverted signal with the delayed signal for positive pulse reading and the inverse of this for reading negative pulses. Besides providing sharply defined outputs, this method permits reading both positive and negative pulses from a single channel without interference or ambiguity. Examples are given for using this facility for checking purposes and for storage of two types of data in a common channel.

A Digital Voltage Encoder—J. R. Zweizig

A two-channel voltage encoder having a sampling rate of 40 numbers/second in each channel has been designed for use in a data reduction system. The data are recorded on single-channel magnetic tape in the form of 10-digit binary numbers with an accuracy of ± 0.1 per cent. The range of input voltage is 0 to 1 volt and may be lowered to 0 to 10 mv through the use of chopper amplifiers. This lower range of input voltage covers the voltages generated by analog transducers commonly employed to measure temperatures, pressures, flow rates, and thrusts.

A New Method of Generating Functions—Lazarus G. Polimerou

As a result of a pressing need for function generators, a new method of function generation has been developed. The underlying principle of this function generator is the application of ordinary pulse techniques in such a way as to produce a function. The simplicity of design, the high accuracy attainable, the simple type of construction are the outstanding features of this general-purpose function generator.

In order to compare this new type of function generator with those presently being used, three important types are discussed. These generators are of the general-purpose, electric and photoelectric types; other comparable electromechanical types are excluded.

A Function Generator for the Solution of Engineering Design Problems—C. J. Savant and R. C. Howard

The solution of nonlinear engineering design problems demonstrates the need for a special function generator. The generator described in this paper satisfies this need. The basic components of the unit are discussed and the forms of functions which can be generated are shown. Accuracy is estimated by comparison of an oscillogram with the calculated curves. It is concluded from tests on the system that the function generator is a valuable aid in the handling of nonlinear design problems.

News

Contributors

Review Section

INFORMATION THEORY

PGIT-4, SEPTEMBER, 1954

(Symposium on Information Theory, M.I.T., Cambridge, Mass., Sept. 15–17, 1954)

Preface—W. G. Tuller

A New Basic Theorem of Information Theory—A. Feinstein

A new theorem for noisy channels, similar to Shannon's in its general statement but giving sharper results, is first formulated and proven. It is then shown that the equivocation of the channel defined by the theorem vanishes with increasing code length. The remaining sections are devoted to various generalizations; in particular, to defining a continuous channel in a manner that permits the application to it of the results given above. The detailed proof of the equivalence of this definition and Shannon's is given in an appendix.

Binary Coding—M. J. E. Golay

It is shown that efficient binary symbol coding with 2-error corrections is impossible, and, more generally, that a search for efficient e-error symbol coding need only be a finite one, as an upper bound exists beyond which demonstrable impossibility sets in.

The question remains open whether there exists more than one case of efficient binary message coding with more than one error correction, and the possibly greater fruitfulness of this approach is suggested by a few cases of inefficient coding where more information can be transmitted by message coding than by symbol coding.

Error-Free Coding—P. Elias

Some simple constructive procedures are given for coding sequences of symbols to be transmitted over noisy channels. A message encoded by such a process transmits a positive amount of information over the channel, with an error probability which the receiver may set to be as small as it pleases, without consulting the transmitter. The amount of information transmitted is less than the channel capacity, so the procedures are not ideal, but they are quite efficient for small error probabilities.

A Class of Multiple-Error-Correcting Codes and the Decoding Scheme—I. S. Reed

A procedure for constructing one-error-correcting and two-error-detecting systematic codes has been introduced by R. W. Hamming. Some examples of n-error-correcting and $(n+1)$ error-detecting systematic codes, for the cases where both the code length and $(n+1)$ are powers of two, are presented. The decoding scheme in this report differs from Hamming's scheme in that the encoded message will be extracted directly from the possibly corrupted received code by a majority testing of the redundant relations within the code.

Coding for Constant-Data-Rate Systems—R. A. Silverman and M. Balser

This paper discusses the use of error-correcting codes in reducing the error rate in communication systems which transmit data at a constant rate. A new single-error-correcting code (The Wagner code) is described and analyzed. Its performance is compared with that of Hamming's single-error-correcting code, and is found to be superior for many communication applications. The principle of the Wagner code is then used to construct two new multiple-error-correcting codes, whose performance is compared with that of Reed's multiple-error-correcting code. If the criterion of performance is that the frequency of errors in sequences of m binary digits be as small as possible, it is found that each of the multiple-error-correcting codes is especially suited to a certain range of values of m.

Information, Organization and Systems—J. Rothstein

The object of this paper is to develop and apply a mathematical concept of organization and of systems. It is very closely related to the information concept and provides the link whereby the theorems of communication theory become generalized and applicable to systems in general. Brief applications are given to system reliability, the significance of organization theory for circuit design, and production and quality control from a system's viewpoint.

An Information-Theoretical Model of Organizations—M. Kochen

The description of certain organizational systems is axiomatically formalized. Such systems are regarded, relative to an outside observer, as groups of participants, capable of selecting from a set of alternatives, such as to maximize the value to themselves, according to

a subjective scale; each member is able to store, subject to limited storage capacity, the communicated choices from one or more others, in suitably coded form.

The only data assumed to be available to each participant is a matrix of the encoded choices of others, and the value accruing from the combination, for a sample of several time periods. In terms of this, expressions for efficiency and order of organization are obtained which agree with established results for large samples. Communication patterns, application of Shannon's fundamental theorem, the temporal behavior of systems in special cases, and some illustrative examples and applications are studied.

Simulation of Self-Organizing Systems by Digital Computer—B. G. Farley and W. A. Clark

A general discussion of ideas and definitions relating to self-organizing systems and their synthesis is given, together with remarks concerning their simulation by digital computer. Synthesis and simulation of an actual system is then described. This system, initially organized at random within wide limits, organizes itself to perform a simple prescribed task.

A Study of Ergodicity and Redundancy Based on Intersymbol Correlation of Finite Range—S. Watanabe

Some of the basic concepts of information theory are critically reviewed in the light of a generalized formulation of the theory of Markoff's chains, in which the initial and final states are sequences of symbols of different lengths, and occurrence of symbols is governed by intersymbol correlation probability of finite range. In particular, the conditions of ergodicity and the structure of "ergodic subsets" of sequences of arbitrary length are carefully discussed. A mathematical method is developed to determine the "range" and "strength" of intersymbol correlation. A brief summary of the content is given at the end of Section 1.

Multivariate Information Transmission—W. J. McGill

A multivariate analysis based on transmitted information is presented. It is shown that sample transmitted information provides a simple method for measuring and testing association in multidimensional contingency tables. Relations, with analysis, of variance are pointed out, and statistical tests are described.

Choice and Coding in Information Retrieval Systems—C. N. Mooers

Information retrieval machines are devices for indexing and selecting information in a library. The operation of these machines is based upon an arbitrary code system, in terms of which the machines' operations are defined. This paper applies the formalism of communication theory as developed for signalling to machines for information retrieval. Three topics are discussed in this paper. 1) The retrieval system analog to $H = -\Sigma p_i \log p_i$, the measure of the output of a source, is developed. 2) The retrieval system analogs to synchronous and asynchronous multiplex types of coding are described and channel capacities are discussed. For the latter type of coding, called "superimposed," a new limit on "channel capacity" of $\log_e 2$ bits per site is given. 3) Selection errors due to coding are discussed, and it is shown that the frequency of errors can be made arbitrarily small for the superimposed type of coding in analogue to the result for signalling.

Modern Statistical Approaches to Reception in Communication Theory—D. Van Meter and D. Middleton

When reception in the theory of communication is recognized as a problem in statistisal inference, system design and system analysis appear as the counterparts of designing and evaluating statistical tests. This paper discusses the optimum properties of designs based on statistical decision theory from the risk point of view, and from that of information theory. Connections between risk and information loss are established, which result in a unified theory of system design. This includes Minimax methods capable, in principle, of handling all degrees of *a priori* knowledge of signal and noise statistics, new methods for comparing actual and ideal systems for the same purpose, and new interpretations of previously used formulations as special cases of the more general theory. Both detection and extraction of signals in noise are considered, the former as a problem of testing statistical hypotheses and the latter as one of estimating parameters.

A Non-Linear Prediction Theory—R. F. Drenick

The paper deals with the smoothing and the prediction of certain signals in noise. It is, more particularly, a study of the conditions under which the optimum sampling filters obtained from that theory are non-linear, and what improvement in performance can be expected. The theory is restricted to the case of signals which are representable over a reasonable period by a polynomial in time. It is found that nonlinear prediction filters result, in general, when the noise is non-Gaussian, regardless of error criterion. Rules are established for the synthesis of such filters. The performance is calculated for a specific case and indicates considerable improvement.

The Detection of Signals Perturbed by Scatter and Noise—R. Price

The functional form of a probability computing receiver is determined for the case of signals perturbed both by transmission through a "scatter" channel and by the addition of Gaussian noise. The "scatter" channel considered here takes the form of a complex, multiplicative random process. In general, the receiver computations involve the operations of linear transformation and matrix inversion. However, in the case of small signal-to-noise ratios, a considerable simplification results and a practical receiver structure is obtained.

The Theory of Signal Detectability—W. W. Peterson, T. G. Birdsall and W. C. Fox.

The problem of signal detectability treated in this paper is the following: suppose an observer is given a voltage, varying with time during a prescribed observation interval, and is asked to decide whether its source is noise or signal plus noise. What method should the observer use to make this decision, and what receiver is a realization of that method? After giving a discussion of theoretical aspects of this problem, the paper presents specific derivations of the optimum receiver for a number of cases of practical interest.

The Human Use of Information. I: Signal Detection for the Case of the Signal Known Exactly—W. P. Tanner, Jr., and J. A. Swets

A theory of visual detection is developed, based on the model provided by the theory of signal detectability, and, more generally, by the theory of statistical decision. Two experiments are reported which test some predictions of the theory for the case of the signal-known-exactly. These experiments demonstrate that the human observer tends toward optimum behavior, where optimum behavior is defined as that behavior which maximizes the expected gain from the decision. Their results show the proportion of correct detections to be dependent upon the proportion of false alarms; they indicate that neural activity is a power function of signal intensity. The data also demand a re-evaluation of the threshold concept. Predictions are made for the data obtained, using two different methods of response, forced-choice and yes-no, and the internal consistency of the theory is demonstrated. The predictions of the theory are compared with contrasting predictions of conventional sensory theory; the data are also related to conventional theory.

The Human Use of Information. II: Signal Detection for the Case of an Unknown Signal Parameter—W. P. Tanner, Jr., and R. Z. Norman

Two specific cases of signal detection, involving uncertainty in the frequency of a sound signal, are compared with the case of the signal-known-exactly. In the first case, the signal is either of two known frequencies; in the second case, the signal is any frequency within a given range. It is suggested that detection behavior that is optimal for the three cases requires a dual mechanism; a combination of a wide-open receiver and a panoramic receiver. Evidence is presented that supports the existence of such a mechanism. Estimates of the bandwidth and scan-rate of the receiver are included.

RADIO TELEMETRY & REMOTE CONTROL

PGRTRC-1, August, 1954

Editorial

The Integrated Air Force Missile Test Center Data Processing Facility—C. F. West

Basic Design of Commutating Devices—M. V. Kiebert, Jr.

High Speed Digital Computers—Marc Shiowitz

ULTRASONICS ENGINEERING

PGUE-2, November, 1954

Mechanical Impedance Transformers in Relation to Ultrasonic Machining—Lewis Balamuth

Ultrasonic machining of hard materials (Process invented by Balamuth. U. S. Patent No. 2,580,716), such as cemented carbides, steel, sapphire, etc., requires the use of an acoustical transmission line. This line generally consists of a specially shaped solid body designed so as to convey the ultrasonic vibrations of the transducer to the tool which operates on the work piece. The paper deals with a general type of mechanical impedance transformer suitable for the above purposes. The theory has been cast in such form that the design of such lines may be made from universal equations and graphs. Because the equations appear in terms of ratios, one set of curves applies to all materials and to any desired frequency range. As is the case generally with transformers, the systems described in this paper have applications in many cases where transformation of the ultrasonic vibration amplitude is desired in going from the source generator to the output end of the system.

Ultrasonic Machining of Tungsten Carbide—Deiter Goetz

The results of an experiment are given in which the ultrasonic machining velocity into tungsten carbide is related to changes in the tool tip material and to changes in the thickness of the abrasive slurry. Some mention of the experimental technique is made, and control of the significant variables is discussed.

An Ultrasonic Machine Tool—Neil Clark, Jr.

The relatively new process of ultrasonic grinding is discussed in general, and a brief description given of a complete ultrasonic machine tool equipment. Operation of the associated magnetostrictive transducer is given in some detail with reference to simple electrical analogies. Design objectives, with reference to machine operation, are explained and a brief analysis given of mechanical transformers as they pertain to coupling the maximum of energy to the cutting load. A description of the 25 kc electronic power driver is given with a method of coupling it to the transducer. Some remarks on theories of material removal and application of the equipment are given in conclusion.

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Books

Bibliography on Research Administration by George P. Bush

Published (1954) by the University Press of Washington, D. C., Washington 4, D. C. 128 pages+18 page index+iv pages. 6¼×9¼. $4.00.

G. P. Bush is a professor at The American University, Washington, D. C.

This volume is a collection of exceedingly brief abstracts in the field of research administration and management, systematically arranged and suitably indexed. The editing of material into chapters dealing with "Research Process," "Budget and Finance," "Organization and Management," "Personnel Administration," "External Relations" and "Research in Action" should assist the research director, project head, or serious research student.

This is convenient reference material but timely only at the moment. In this connection, no closing date for inclusion of material is mentioned although the statement is made that "eleven hundred references are suggested . . . chosen on a selective basis and primarily dated since 1945." Since the management papers presented at the 1953 National Electronics Conference are omitted, one draws the conclusion that the volume went to press prior to the close of 1953.

Review of current periodical literature as well as the usual texts adds much to the value of the book. It is in the former medium that one notes and follows the evolution of many new ideas in management direction.

Ralph I. Cole
Melpar, Inc.
Alexandria, Va.

Transistoren by M. J. O. Strutt

Published (1954) by S. Hirzel Verlag, Claridenhof, Gotthardstrasse 6, Zurich 2, Switzerland. 153 pages +4 page index+9 page bibliography. Illustrated. 6×9.

M. J. O. Strutt is Professor and Chairman of the Department of Advanced Electrotechnics at the Swiss Federal Institute of Technology at Zurich.

This well organized book represents the continental viewpoint of a substantially American development in an excellent manner for the American reader. It is a highly commendable textbook in German for European consumption.

The scope indicated by the subtitle of the book, "Operation, Properties and Application," is well covered. According to the foreword, Dr. Strutt's Institute has conducted research since 1948 in the field of transistors and this seems to be indicated by the treatment of the subject, contrary to some recent books which betray a solely reading knowledge of the subject.

The introductory chapter treats such questions as: "What is a Transistor," "Transistor Curves," "The Comparison of These Curves with Resistance Four Pole Networks," thus introducing the student directly to the subject but guiding him through familiar ground. Subsequent chapters treat electrons in materials (contrasted with electrons in vacuum), electronics of contacts, the application of contact transistors and junction transistors. This part of the book is written in a simple but scholarly manner, with an emphasis on teaching engineers rather than physicists.

The latter part of the book, covering transistor circuitry, seems to benefit most from the actual work carried on by Dr. Strutt and his students. These latter chapters treat "Duality and Analogy of Circuits," "Transistors as Small Signal Amplifiers," "Transistors in Power Output Stages," "Oscillators," and "Test Methods and Measurements."

The book contains a bibliography of 148 titles, mainly from American sources, which is quite gratifying.

George C. Sziklai
RCA Labs.
Princeton, N. J.

Automatic Digital Calculators by A. D. Booth and K. H. V. Booth

Published (1953) by Academic Press, Inc., 125 E. 23rd Street, New York 10, N. Y. 216 pages+5 page index+10 page bibliography+vii pages. 91 figures. 3 plates. 5½×8½. $6.00.

This book will introduce the novice to almost every part of the digital computing field. The first three chapters review the history of the field and the major machines built prior to 1953. These chapters are as easy to read as a novel. Following chapters present detailed logical designs for control systems, arithmetic units and other major parts of computers. The casual reader may have difficulty with this material, but sufficient text accompanies the block diagrams for a determined reader to master the material. Later chapters discuss building blocks based on vacuum-tubes, semiconductor diodes and magnetic elements. Most storage devices are considered in detail adequate for the beginning student. The book is not limited to engineering and has considerable information on programming. The reader can learn, for example, ways of including a sub-routine in a master-routine, of coding decimal to binary conversion, and programming interpolation in tables. Remarks on computer applications are given point by the consideration of a problem in X-ray crystal structure analysis. The discussion of the application of machines to games and translation summarizes our incomplete understanding of those matters.

The book is a valuable contribution to education in the computer field. The isolated student will find it possible to study the book and obtain a fair competence in computer engineering and programming.

J. H. Felker
Bell Telephone Labs.
Whippany, New Jersey

Abstracts and References

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research, London, England, and Published by Arrangement with that Department and the *Wireless Engineer*, London, England

NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

Acoustics and Audio Frequencies . . 1717
Antennas and Transmission Lines . . 1718
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The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number and is not to be confused with the Decimal Classification used by the United States National Bureau of Standards. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

A new section has been introduced covering the general technique of electromagnetic waves, oscillations and pulses (i.e. transmission lines and circuits), as distinct from specific applications to telecommunications. The new section is numbered 621.37, with subdivisions. Full details of the new classification, and of the numbers which become obsolete as a result of its introduction, are given in PE Note 535, obtainable from The International Federation for Documentation, Willem Witsenplein 6, The Hague, Netherlands, or from The British Standards Institution, 2 Park Street, London, W.1., England.

Section 621.396.67, dealing with Antennas, has been modified and expanded; details of the new classifications are given in PE Note 519.

Section 621.396.96, with subdivisions, has been introduced to cover Radar; details of the new classifications are given in PE Note 518.

New Subject Section

A section headed Automatic Computers has been introduced.

The Index to the Abstracts and References published in the PROC. I.R.E. from February 1953 through January 1954 is published by *Wireless Engineer* and included in the March 1954 issue of that journal. Copies of this issue may be purchased for $1 (including postage) from the Institute of Radio Engineers, 1 East 79th Street, New York 21, N.Y. As supplies are limited, the publishers ask us to stress the need for early application for copies. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

ACOUSTICS AND AUDIO FREQUENCIES

534.232 2829

Determination of Equivalent-Circuit Constants of Piezoelectric Resonators of Moderately Low Q by Absolute-Admittance Measurements—G. E. Martin. (*Jour. Acous. Soc. Amer.*, vol. 26, pp. 413–420; May 1954.) Previously established analysis [3164 of 1953 (Cady)] is extended to apply to the conditions governing underwater sound transducers or $BaTiO_3$ resonators in air. The errors inherent in the derived formulas are evaluated by exact numerical calculations; they are very much smaller than those resulting with the conventional method of calculation.

534.232 2830

The Electro-Acoustic Sensitivity of Cylindrical Ceramic Tubes—R. A. Langevin. (*Jour. Acous. Soc. Amer.*, vol. 26, pp. 421–427; May, 1954.) An exact formula is derived for the sensitivity of cylindrical $BaTiO_3$ transducers for underwater use, for various boundary and polarization conditions.

534.26+[538.566:535.42 2831

Diffraction by Two Parallel Planes of Finite Length—Williams. (*See* 2908.)

534.321:534.75 2832

Subjective Musical Pitch—W. D. Ward. (*Jour. Acous. Soc. Amer.*, vol. 26, pp. 369–380; May, 1954.) A distinction is drawn between pitch and frequency analogous to that between loudness and intensity. Experiments are described which demonstrate the validity of the subjective measure "octave" for determining scales for pure tones presented successively. Day-to-day variations of judgments are ascribed to a binaural effect.

534.321.9 2833

Ultrasonic Spectrum of Cavitation Noise in Water—R. H. Mellen. (*Jour. Acous. Soc. Amer.*, vol. 26, pp. 356–360; May, 1954.) An investigation of the noise spectrum produced by transient cavities of the type encountered in tip cavitation of marine propellers. Measurements over the frequency range 1–3,500 kc are reported.

534.321.9:534.613 2834

Radiation Pressure and the Nonisotropic Nature of Acoustic Pressure at High Intensity—V. Gavreau and M. Miane. [*Compt. Rend. Acad. Sci. (Paris)*, vol. 238, pp. 2148–2150; May 31, 1954.] Using a plane wave of frequency 77.1 kc, particle velocity and sound intensity were determined from measurements of the amplitude of vibration of the generator. Radiation pressure was measured with a sensitive balance, using conical obstacles with the apex directed towards the source, and was found to vary as the square of the cosine of the cone base angle. Cones with various types of surface, e.g. smooth, absorbent, stepped, perforated, were used. The results indicate that for intense radiation the pressure is greatest in the direction of propagation; the effect may be accompanied by an increase in the velocity of propagation.

534.6 2835

Standard Test Gear developed by the Acoustics Department of C.N.E.T.—P. Chavasse and P. Riéty. (*Ann. Télécommun.*, vol. 9, pp. 58–71 and 94–98; March/April, 1954.) Tests are carried out by the Department on (a) electro-acoustic transducers, (b) acousto-electric transducers, (c) acoustic networks, e.g., hearing aids and acoustic transmission systems, (d) af amplifiers and oscillators, (e) sound sources. The equipment developed for test purposes is briefly described, but details are given of a special cro deflection circuit and of the synchronization of the frequency-sweep oscillator and the recording gear. Equipment for work in physiological acoustics is also provided.

534.6 2836

Determination of Absolute Sound Levels and Acoustic Absorption Coefficients by Thermocouple Probes—W. J. Fry and R. B. Fry. (*Jour. Acous. Soc. Amer.*, vol. 26, pp. 294–317; May, 1954.) Theoretical and experimental aspects of the method are discussed. The sound intensity is determined from the specific heat and the absorption coefficient of the material in which the thermocouple is embedded and the observed rate of change of temperature at the thermocouple junction on application of a sound pulse.

534.75 2837

Responding to One of Two Simultaneous Messages—W. Spieth, J. F. Curtis, and J. C. Webster. (*Jour. Acous. Soc. Amer.*, vol. 26, pp. 391–396; May, 1954.) The effect of the following factors on the ability of an operator (e.g. an air-control-tower operator) to attend to one of two simultaneous messages was investigated: (a) horizontal separation between message sources, (b) cue lamps, (c) a "pull-down" arrangement for switching a chosen message from the originating loudspeaker into one near his left ear, (d) use of different af filters in the different channels. Suitable adjustment of factors (a) and (d) improved the operator's performance greatly.

534.75 2838

Responding to Both of Two Overlapping Messages—J. C. Webster and P. O. Thompson. (*Jour. Acous. Soc. Amer.*, vol. 26, pp. 396–402; May, 1954.) Experiments are described which show that the ability of an operator to respond to overlapping messages decreases as the overlap time increases. The leading message is understood better unless its intensity is considerably lower. An examination is made of factors in the messages affecting ease of differentiation.

534.78 2839

On the Identification of Speakers by Voice—I. Pollack, J. M. Pickett, and W. H. Sumby. (*Jour. Acous. Soc. Amer.*, vol. 26, pp. 403–406; May, 1954.)

534.78 2840

Design of Visible Speech Devices—G. E. Peterson. (*Jour. Acous. Soc. Amer.*, vol. 26, pp. 406–413; May, 1954.)

534.79 2841

Loudness Relations in Two-Component Tones—A. M. Small, Jr. and W. R. Thurlow. (*Jour. Acous. Soc. Amer.*, vol. 26, pp. 381–388; May, 1954.) Experiments were conducted to determine how the loudness of each component of a two-tone stimulus changes as the intensity of the higher-frequency component is raised from masked threshold to well above that level. Results are shown graphically for various pairs of frequencies. Opposite loudness-variation effects were observed for higher-frequency components above and below 1 kc respectively.

534.833 2842

Sound Insulation with Light-Weight Structures—J. Pujolle and R. Lamoral. (*Onde élect.*, vol. 34, pp. 384–386; April, 1954.) Measurements are reported on structures of weights up to 15 kg/m², suitable for movable partitions or soundproof doors. With complex structures based on duralumin and compressed wood, results appreciably better than indicated by the mass law were obtained. A new method of fixing, using a multicellular plastic, gave good insulation and damped out low-frequency vibrations.

534.833.4 2843

Sound Absorption by Perforated Porous Tiles: Part 1—U. Ingard. (*Jour. Acous. Soc. Amer.*, vol. 26, pp. 289–293; May, 1954.) Formulas are derived from which the acoustic properties of the tile can be calculated in terms of its geometry and the physical properties of the material. Measurements of the normal impedance as a function of perforation depth are in good agreement with calculated values.

534.845 2844

Acoustic Factors in the Design of Noise-Reducing Telephone Booths—D. E. Bishop, F. G. Hewitt, and D. B. Callaway. (*Jour. Acous. Soc. Amer.*, vol. 26, pp. 318–322; May, 1954.) Investigations are described resulting in a triangular design for a doorless booth.

621.395.625.3 2845

Avoiding Echo Phenomena in Multiple-Track Magnetic Recording Systems—G. Ouvrier. (*Funk u. Ton*, vol. 8, pp. 174–178; April, 1954). Results of experiments show that crosstalk between neighboring sound tracks originates in the reproduction and not during recording. The dependence of the effect on recorded frequency, running speed, and track spacing is considered.

621.395.625.3:538.221 2846

Ferrites as Core Materials for Magnetic Recording Heads—O. Schmidbauer. (*Funk u. Ton*, vol. 8, pp. 169–173; April, 1954.) An expression is derived for the useful induced flux, taking account of the "magnetic roughness" of the ferrite surface due to the coarse granular structure. Roughness values for different ferrites are determined by comparison with a mumetal head. Ferrite heads should be suitable where medium and high tape speeds are used, and in conjunction with magnetic films.

621.395.625.3:621.397.5 2847

Recording Television and Sound Signals on Magnetic Tape—(*Elec. Eng.*, vol. 26, pp. 292–295; July, 1954.) See 2517 of August (Olson et al.).

621.395.625.6:778.534.4 2848

A Photographic Technique of Sound Recording on Glass Discs—A. J. Forty. (*P.O. Elec. Eng. Jour.*, vol. 47, part 1, pp. 19–24; April, 1954.) A description of the process used by the British Post Office, with discussion of factors limiting performance. The main application is in announcing machines for the telephone service.

ANTENNAS AND TRANSMISSION LINES

621.372.2+621.317.34.029.64 2849

An Experimental Investigation of the Single-Wire Transmission Line—T. E. Roberts, Jr. (*Trans. I.R.E.*, vol. AP-2, pp. 46–56; April, 1954.) The lines investigated were dielectric-coated wires of brass and hard-drawn copper of length about 80λ. Measurements were made at λ 3.2 cm to determine (a) charge and current distributions, (b) attenuation and phase velocity, and (c) the radiation conductance applicable in the equivalent circuit for the transition from a flanged coaxial line to a single-wire line. Measurement techniques are described. Results verify the theoretical result that the single-wire line, if sufficiently long, may be treated as a simple transmission line (1580 of 1953). Comparison with data for uncoated wires clearly shows the guiding properties of a thin dielectric coating. The efficiency of horn couplings used in the measurements was also determined experimentally.

621.372.2+621.372.8 2850

Use of Short-Circuiting Pistons in the Study of Junctions and Directional Couplers—A. Weissfloch. (*Ann. Télécommun.*, vol. 9, pp. 81–92; March, 1954.) Line junctions are treated as six- or eight-pole networks, according as they are of *T* or *X* type. For junctions for particular applications, a method of measurement using nonreflecting terminations is often satisfactory. For six-pole junctions for general use a more accurate method is described in which the output lines are terminated by short-circuiting pistons and the only measurement required is the position of the voltage nodes on the input line. Proof of the theory of the method and notes on experimental technique and the accuracy attainable are presented. The application of the method to junctions of eight-terminal type and to directional couplers is also explained.

621.372.2 2851

Waves of Current without Voltage and of Voltage without Current in Two Parallel Lines—M. Cotte. (*Onde élect.*, vol. 34, pp. 381–383; April, 1954.) Analysis based on Laplace transforms is given. See also 3776 of 1947, where analysis is based on Carson transforms.

621.372.2:621.315.212 2852

Application of Brewster's Angle to the Design of Coaxial-Line Components for Microwaves—B. A. Dahlman. (*RCA Rev.*, vol. 15, pp. 238–251; June, 1954.) A theoretical investigation is made of the conditions under which refraction without reflection can be obtained at the boundary between two dielectrics, with particular reference to the use of conical lines as adapters between coaxial cylindrical systems. The observed vswr for an experimental system using such an adapter between two 50-Ω lines of over-all diameter $\frac{7}{8}$ inch and $\frac{3}{8}$ inch respectively was <1.06 over the frequency range 2–5.8 kmc.

621.372.8+621.396.677.7 2853

On the Theory of Corrugated Plane Surfaces—R. S. Elliott. (*Trans. I.R.E.*, vol. AP-2, pp. 71–81; April, 1954.) Analysis is based on applying Floquet's theorem to Lucke's resonator principle for determination of the propagation characteristics of a corrugated surface. If the noncorrugated side and top walls of a waveguide with corrugated bottom wall are gradually flared out and terminated, a surface wave can be satisfactorily launched on the extended bottom wall; if at the other end the depth of corrugations gradually decreases to zero, a good match to a rectangular waveguide may be achieved, the system operating efficiently with TE_{01}-mode excitation. Approximate radiation patterns are derived for such a system terminated by (a) an infinite, and (b) a finite, ground plane. For the latter case approximate theory gives an upper limit for the tilt angle of the main beam and a lower limit for beam width. The suppression of feed radiation and the elimination of higher-order modes are considered. Experimental data fit the theory reasonably well.

621.372.8 2854

Circumferential Gap in a Circular Waveguide Excited by a Dominant Circular-Electric Wave—L. S. Sheingold and J. E. Storer. (*Jour. Appl. Phys.*, vol. 25, pp. 545–552; May, 1954.) The gap is considered as a free space coupling two semi-infinite waveguides, the width of the gap being small compared with the guide wavelength. The effect of the gap on the propagation of the TE_{01} mode is examined, the scattering matrix being derived by a variational method. Results of measurements on a brass waveguide of internal diameter 5.755 inches, at a free-space wavelength of 10 cm, support the theory.

621.372.8 2855

Propagation of Microwaves through a Cylindrical Metallic Guide Filled Coaxially with Two Different Dielectrics: Part 5—S. K. Chatterjee. (*Jour. Indian Inst. Sci.*, section B, vol. 36, pp. 48–58; April, 1954.) General conditional equations are formulated from which the propagation characteristics of any hybrid mode can be derived using the theory presented in Part 4 (1992 of July). Calculations are made for the special case of a simple dielectric rod. This behaves as a guide, or as an antenna, accordingly, as its radius is under or over a critical value depending on the mode of excitation.

621.372.8 2856

Attenuation Coefficient of Waveguides loaded with Circular Irises—R. Combe. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 2063–2065; May 24, 1954.] An approximate expression is obtained for the attenuation of a rotationally symmetrical mode with a longitudinal electrical field component in a loaded circular waveguide. The formulas obtained are useful for the design of electron accelerators.

621.372.8:621.385.029.64 2857

Broadbanding of Resonant-Type Microwave Output Windows—T. S. Chen. (*RCA Rev.*, vol. 15, pp. 204–229; June, 1954.) The bandwidth of magnetron output windows of the type described previously (2307 of August) is increased by placing a resonant iris at a distance of λ/4 on each side of the ceramic plate. Bandwidths of 1.0 and 1.57 kmc, centered at 9.15 and 9.35 kmc respectively, have been obtained for two different ceramics. The vswr was <1.1 in both cases.

621.396.67 2858

Skeleton Slot Aerial—H. B. Dent. (*Wireless World*, vol. 60, pp. 399–401; August, 1954.) An antenna which has operated satisfactorily on the 2-m amateur band has the form of a rectangular metal frame of length slightly less than λ/2 and width about a third of the length. When erected with the long side vertical, its radiation is horizontally polarized, and it has a useful power gain over a simple horizontal λ2 dipole. Methods of combining with a reflector are discussed. Measured polar diagrams are shown.

621.396.67 2859

The Theory and Application of the Radiation Mutual-Coupling Factor—E. A. Blasi. (Proc. I.R.E., vol. 42, pp. 1179–1183; July, 1954.) A formula is derived for the mutual impedance of two radiating elements in terms of the self-impedances. A convenient new parameter, the radiation mutual-coupling factor, is defined which can be accurately measured. The technique is illustrated for the case of (a) a pair of identical λ/4 antennas, and (b) a pair of identical axial-mode helical radiators. Experimental results support the theory.

621.396.673 2860
Impedance of a Top-Loaded Antenna of Arbitrary Length over a Circular Grounded Screen—J. R. Wait and W. J. Surtees. (*Jour. Appl Phys.*, vol. 25, pp 553-555; May, 1954.) Abbott's method (2729 of 1952) is used to calculate the approximate self-impedance of a thin top-loaded vertical antenna over an infinitely conducting screen lying on a finitely conducting ground.

621.396.677:621.396.721.029.51/53 2861
The Problem of the Long-Wave Radio Relay—Kümmich. (*See* 3042.)

621.396.677.7 2862
On the Bandwidth of Cavity Antennas—M. H. Cohen. (*Jour. Appl Phys.*, vol. 25, pp. 582–587; May, 1954.) To simplify calculation, a cavity is considered which comprises a shorted square waveguide, of side a, opening into a half-space; the resonance frequency and the bandwidth ($1/Q$) are determined from transmission-line theory. The analysis depends on an equivalent aperture admittance. The minimum value of Q for such an antenna is given by $0.424\,(a/\lambda)^{-3}$ for values of $(a/\lambda)<0.35$. The cavity depth required to obtain minimum Q is of the order of $4a$. The required dielectric constant increases rapidly as a/λ decreases. Experimental results which support the theory are presented.

621.396.677.7 2863
Theory of Thin Aerials in Endovibrators—A. V. Gaponov and M. L. Levin. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 95, pp. 1193–1196; April 21, 1954. In Russian.] A method is developed of solving the problem of excitation by currents flowing in the perfectly conducting walls of an endovibrator with thin antennas, slots, or both. The problem of excitation at the resonance frequency has been considered by Fel'd in his book "Foundations of the Theory of Slot Aerials" (in Russian), Moscow, 1948.

621.396.677.8 2864
Propagation of Microwaves between a Parallel Pair of Doubly Curved Conducting Surfaces—K. S. Kunz. (*Jour. Appl. Phys.*, vol. 25, pp. 642–653; May, 1954.) Mathematical theory is given for microwave lenses comprising parallel plates with the intervening space filled with a substance of arbitrary refractive index. The spacing between the plates is assumed small compared to λ; only the TEM mode is then propagated. A family of equivalent lenses is shown to exist; the conditions are determined for one member of the family to have unity refractive index everywhere. It is shown that the Rinehart analog of the Luneberg lens and the RCA "R-2R" lens are members of a family with analogous focusing properties.

621.396.677.83 2865
Wide-Band Aerials with High Gain Efficiency—H. Meinke. (*Fernmeldetech. Z.*, vol. 7, pp. 161–170; April, 1954.) Antennas for the frequency range 200 mc-2kmc, using parabolic-cylinder reflectors with conducting ends, are discussed in relation to the requirements of high gain in the main radiation direction, low side lobes, and satisfactory wide-band matching. It is shown that when the dimensions of the reflector are not greater than a few wavelengths, efficiencies of over 100 per cent can be attained in theory and of nearly 100 per cent in practice. Radiation pattern and matching are discussed in relation to problems of construction. The reflector may be combined with a horn.

621.396.677.831 2866
An Analysis of Passive Reflector Antenna Systems—R. E. Greenquist and A. J. Orlando. (Proc. I.R.E., vol. 42, pp. 1173–1178; July, 1954.) Equivalent-aperture theory is used to derive the gain and radiation pattern for a system with a plane rectangular reflector at 45 degrees to the antenna beam. Experimental results support the method of analysis. Side-lobe levels and mutual coupling are also discussed

621.396.677.833.1 2867
Field Problem for a Parabolic Reflector—G. A. Grinberg, N. N. Lebedev, I. P. Skal'skaya, and Ya. S. Uflyand. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 95, pp. 961–963; April 11, 1954. In Russian.] The field due to a linear antenna placed along the focal line of a parabolic cylindrical reflector is considered. The solution of the equation $\Delta u+k^2u=0$, where u is the complex amplitude of the electric field, is given as the sum of two terms, the first involving the Hankel function of the second kind of kr, where k is the wave number and r the distance, the second involving a definite integral of hypergeometric functions of a parameter ν. For $|\alpha|<\beta$, where α and β are the parabolic coordinates, the integral can be evaluated in the form of an infinite series. Magnus's solution (1966 of 1942) is discussed and found to be valid in this region and hence, from the properties of elliptic integrals, in the whole region where u is a regular function.

AUTOMATIC COMPUTERS

681.142 2868
A Versatile Electronic Function Generator—R. Tomovich. (*Jour. Frank. Inst.*, vol. 257, pp. 109–120; February, 1954.) Two types of electronic circuit for performing any desired transformation on two arbitrary voltage waveforms are described. The general characteristics and accuracy of such devices when used in repetitive differential analyzers are discussed.

CIRCUITS AND CIRCUIT ELEMENTS

621.318.57:621.314.7 2869
Transistor Shift Registers—R. H. Baker, I. L. Lebow, and R. E. McMahon. (Proc. I.R.E., vol. 42, pp. 1152–1159; July, 1954.) "Three different types of high-speed transistor shift registers are discussed. The high speed (3 to 5 μs per shift pulse) is made possible by the use of nonsaturating bistable circuits. Two general shift registers are described making use of respectively one and two transistors per stage. The third register is of a specialized variety capable of shifting a single digit. In addition, an analysis is made of the triggering requirements of the register using a single transistor per stage."

621.318.57:621.372.8 2870
Waveguide Switching Device using Gas-Filled Tubes—L. Thourel, P. Tcheditch, M. Lamy, and P. Vincent. (*Ann. Radioélect.*, vol. 9, pp. 163–179; April, 1954.) An arrangement for switching a single receiver between two or more channels at a relatively high rate is described. The switching operation is based on detuning an oscillatory circuit consisting of a waveguide section terminated with inductive diaphragms and having a gas-discharge tube located midway between the ends: the hf susceptance of the tube varies with its ionization state, and the tube constitutes an element of a bandpass filter. Analysis is given for the pass band, the switching rate, the electrical inertia, and the noise and insertion loss due to the device. Results obtained with a four-channel system are in good agreement with theory.

621.319.42 2871
Component Design Trends—Fixed Capacitors undergo Miniaturization—F. Rockett. (*Electronics*, vol. 27, pp. 120–125; July, 1954.) First paper of a series describing recent developments. Problems involved in operating at the high temperatures resulting from reduction of the heat-dissipating surface are discussed.

621.372 2872
Some Techniques for Network Synthesis—G. L. Matthaei. (Proc. I.R.E., vol. 42, pp. 1126–1137; July, 1954.) The synthesis of immittances with two poles and two zeros is discussed; Cauer's continued-fraction technique is generalized for use in realizing RLC network immittances. Design procedure is developed for a constant-resistance ladder: this network is shown to have the same field of application as the conventional RLC constant-resistance bridged T, but has the advantages of greater flexibility and less complexity. Complicated transfer functions may be realized in a chain of ladder sections having arm immittances with only two poles and two zeros.

621.372:515.432 2873
Analysis of Network Response to Periodic Waves—P. M. Seal. (*Jour. Frank. Inst.*, vol. 257, pp. 13–24; January, 1954.) The steady-state response of a network to a nonsinusoidal periodic wave may be found by applying the operational-calculus method of Waidelich (1276 of 1946) in a simplified form. The method can be used for any linear network comprising a finite number of lumped elements. A set of rules for obtaining the required inverse transforms together with a table of inverse transforms for square-wave inputs is presented and explained.

621.372:621.314.7 2874
Feedback in Junction Transistor Circuits—D. W. Gade. (*Electronics*, vol. 27, pp. 174–176, 178; July, 1954.) "Series and parallel feedback circuits for a single stage are reduced to a four-parameter equivalent circuit for analysis. Cascaded transistor stages reduced in a similar manner result in either mesh or nodal-derived equivalent circuits."

621.372.011 2875
***Q* as a Mathematical Parameter**—D. Morris. (*Elec. Eng.*, vol. 26, pp. 306–307; July, 1954.) A definition applicable to rc networks and other systems having lumped parameters, and appropriate in cases where Q value is of the order of $\frac{1}{2}$ or less, gives Q in terms of circuit time constant and the differential operator.

621.372.029.3:083.7) 2876
I.R.E. Standards on Audio Techniques: Definitions of Terms, 1954—(Proc I.R.E., vol. 42, pp. 1109–1112; July, 1954.) Standard 54 I.R.E. 3 S1.

621.372.41:621.396.822 2877
Thermodynamical Consideration of Equilibrium Electrical Fluctuations—M. L. Levin. (*Uspekhi fiz. Nauk*, vol. 52, pp. 486–494; March, 1954.) Expressions are derived for the mean energy of electrical fluctuations in an LCR circuit maintained at a given temperature, and for the spectral density of the fluctuation emf. Similar expressions published earlier by Gorelik and Ginzburg are discussed.

621.372.5:512.83 2878
Two Network Theorems concerning Change of Voltage Reference Terminal—J. Shekel. (Proc. I.R.E., vol. 42. p. 1125; July, 1954.) "The first theorem describes the change in the admittance matrix corresponding to the change of the voltage reference terminal ('ground' terminal). The second theorem shows that the determinant of the admittance matrix is invariant under such transformations."

621.372.5:621.316.89:537.311.33 2879
The Application of Some Semiconductors as Logarithmic Elements—N. M. Schaeffer and G. W. Wood. (Proc. I.R.E., vol. 42, pp. 1113–1116; July, 1954.) Voltage dividers incorporating Cu_2O rectifiers to give special characteristics were described by Kallmann (3554 of 1946). For circuits involving high resistances, rectifiers made from other semiconductors may be more suitable. Logarithmic

characteristics obtained using various semiconductor rectifiers are presented. Temperature compensation is achieved by use of negative-temperature-coefficient resistors. The logarithmic response has been obtained at frequencies up to 100 kc.

621.372.54 2880

Practical Low-Pass and Band-Pass Minimum-Phase Networks with Equalized Delay and Aperiodic Transient Response—H. H. Voss. (*Frequenz*, vol. 8, pp. 97–102; April, 1954.) A minimum-phase-shift network with constant delay characteristic is a practical approximation to an idealized low-pass filter, since its attenuation characteristic in the pass band has a similar shape. Conditions for delay equalization are investigated for a unit-step input, in terms of the position of the zeros and poles of the response function. A general method of calculation for filter networks of this type is given. The attenuation and group-delay characteristics in a simple case are compared with those of an equivalent filter based on image-parameter theory.

621.372.54 2881

Audio-Frequency Filter with Continuously Adjustable Attenuation Characteristic—A. Spälti. (*Tech. Mitt. schweiz. Telegr.-Teleph-Verw.*, vol. 32, pp. 121–125; Apri 1, 1954.) A composite filter for the frequency band 85 cps—7 kc is described, comprising two sets each of 24 separately controllable filters with contiguous narrow frequency bands. The controls are arranged so that the positions of the individual knobs give an approximate representation of the over-all frequency characteristic. The device is useful for investigating correcting networks and for producing special effects in broadcasting.

621.372.54 2882

Synthesis of Constant-Time-Delay Networks—M. S. Corrington and R. W. Sonnenfeldt. [*RCA Rev.*, vol. 15, pp. 163–186; June, 1954. *Proc. NEC (Chicago)*, vol. 9, pp. 50–63; 1953.] The synthesis of low-pass filters to have a symmetrical transient response such as is desirable in television receivers is investigated. The desired-frequency response is expanded in a Fourier series; circuit elements corresponding to the terms of the series are constituted by sections of delay lines, which are appropriately combined to give the desired over-all response. Low-pass, bandpass and band-stop filters are considered. Results obtained with experimental filters are discussed.

621.372.54:681.142 2883

Punched-Card Calculating Machines as Aids in the Calculation of Filters—W. Heimann. (*Arch. elekt. Übertragung*, vol. 8, pp. 173–178; April, 1954.) The use of commercial-type punched-card calculators for solving numerical problems involved in insertion-loss filter theory is discussed and illustrated.

621.372.544.2 2884

Comb Filters for Pulsed Radar Use—S. F. George and A. S. Zamanakos. (Proc. I.R.E., vol. 42, pp. 1159–1165; July, 1954.) The use of comb filters for improving signal/noise ratio is discussed in relation to the variation of the frequency spectrum of a radar signal with the number of pulses in the received trains. The relative merits of (a) the simple uniform type, (b) the rectangular $(\sin x/x)^2$ type, and (c) the North type $(\sin x/x)^2$ matched filters are compared. The small gain in sensitivity obtained with type (c) does not warrant the greater complexity of its design.

621.372.632+621.375.23 2885

Gain Stable Mixers and Amplifiers with Current Feedback—G. E. Boggs. (Proc. I.R.E., vol. 42, pp. 1144–1147; July, 1954.) "Narrow-band radio-frequency amplifiers and mixers may be stabilized by negative feedback without increasing the bandwidth excessively. A couple using current feedback is described. This couple requires only a simple resistive beta circuit and may be designed such that the bandpass characteristic is largely independent of the feedback." Experimental results are reported.

621.373 2886

van der Pol's Equation—Z. Akcasu. (*Wireless Eng.*, vol. 31, pp. 198–203; August, 1954.) An analytic method of solving the equation $x''+x=\mu f(x)x'$ is described in which the unknown functions a and θ of the assumed solution $x=a\sin(t+\theta)$ are found by substituting the assumed solution into the integro-differential form of the original equation. The solution of the first approximation thus obtained is identical with solutions previously obtained by others. In the second approximation the method yields the interdependence between amplitude and instantaneous frequency during the transient period. The effect of the harmonic content is calculated. The theory is applied to the case of a tube oscillator with tunedanode circuit.

621.373:621.387 2887

The Ionic Oscillator—T. E. Fairbairn. (*Radio-Electronics*, vol. 24, pp. 76, 80; December, 1953.) A stable device capable of delivering high output at af or rf comprises a simple gas-filled-tube circuit without external or internal resonators. Anode voltage equal to the normal ionization voltage is applied. Am and fm arrangements are indicated. The upper frequency limit is >1.5 mc.

621.373.4 2888

A Reactance Tube Controlled Oscillator of Unusually Wide Frequency Sweep—B. M. Banerjee. (*Indian Jour. Phys.*, vol. 28, pp. 67–74; February, 1954.) Circuits permitting frequency sweeps up to 16 mc are described, one for operation above and the other below 30 mc.

621.373.42 2889

Quartz Generators with Negative Feedback and an Inertial Nonlinearity—M. E. Zhabotinski and D. A. Lisichkin. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 95, pp. 1197–1200; April 21, 1954. In Russian.] Theory is developed on lines similar to that presented by Post and Pit (1860 of 1951). Results show that the best position for the amplitude-limiting element is in a separate negative-feedback circuit of the amplifier part of the generator, since in other positions the element tends to reduce the frequency and phase stability of the oscillator. At frequencies of a few tens of kilocycles stable amplification of the order of 10^5 can be obtained using a three-stage amplifier with phase-correction circuits. Stable amplification at very high and at very low frequencies is also mentioned.

621.373.421 2890

Generalized Equations for RC Phase-Shift Oscillators—S. Sherr. (Proc. I.R.E., vol. 42, pp. 1169–1172; July, 1954.) "General solutions are obtained for three- and four-section phase-shift networks. These solutions are reduced to design equations for a number of specific circuit configurations, and the complete solution is given for a four-section phase-shift oscillator."

621.373.421 2891

Oscillator Feedback Networks of Minimum Attenuation—P. W. Ward. (*Elec. Eng.*, vol. 26, pp. 318–320; July, 1954.) Single-tube oscillators with feedback networks comprising three resistors and three capacitors are discussed. Optimum values for the components are determined by analysis and are plotted as functions of tube output resistance for a given value of tube input resistance. Both shunt-R and shunt-C networks are considered, the former being recommended for very low frequencies and the latter for relatively high ones.

621.373.421.13 2892

Classification of Quartz-Crystal Oscillator Circuits—H. Awender and K. Sann. (*Funk u. Ton*, vol. 8, pp. 202–214 and 253–265; April/May, 1954.) Classification is based on quadripole representation, and two main classes are distinguished according as the circuit is represented by one or two quadripoles in the feedback line. The first class includes π-type networks, subdivided into series-resonance and parallel-resonance circuits, and bridge-type circuits. Variations of the basic circuits are illustrated and formulas are derived for the properties and operating conditions. The main characteristics considered are stabilization factors [see 1274 of 1953 (Kettel)], excitation conditions, operating frequency, and crystal loading.

621 373.52 2893

An Amplitude-Stabilized Transistor Oscillator—E. R. Kretzmer. (Proc. I.R.E., vol. 42, p. 1124; July, 1954.) Correction to paper abstracted in 1352 of May.

621.375.2:621.3.015.7 2894

Effect of Steepness of Rise and Fall of the Input Pulse on the Response of Pulse Amplifiers: Part 1—B. K. Bhattacharyya. (*Indian Jour. Phys.*, vol. 28, pp. 31–47; January, 1954.) Expressions are derived for the response of rc-coupled amplifiers to input pulses of ramp form and of triangular form.

621.375.2.029.4 2895

Some Experiments with Miniature Power Triodes—G. H. Grey. (*Audio*, vol. 38, pp. 19–20, 61; April, 1954.) Description, including complete circuit diagrams, and discussion of performance of two af amplifiers with an output of about 7.5w, one of which uses 6AQ5 tubes in push-pull and the other four 6S4 tubes in parallel in a single-ended output. Using unselected components the latter amplifier gave a more satisfactory performance.

621.375.2.029.4.24 2896

Selective Special Amplifier for a Frequency of 6.25 c/s—K. Bauer and W. Haberditz. (*Arch. tech. Messen*, pp. 89–92; April, 1954.) The design is discussed of a four-stage amplifier for use with a capacitor pickup. Capacitance changes, due to variation of external gas pressure, are $\sim 10^{-6}$—10^{-2}pF in 40 pF. The amplification is sufficient to give an output galvanometer reading for an input of $\sim 2\times 10^{-6}$—3×10^{-4} V across a 10^9-Ω resistance.

621.375.221 2897

Distributed Amplifier for Nuclear Research—K. Enslein. (*Electronics*, vol. 27, pp. 138–141; July, 1954.) Details, including component values, are given of a four-stage amplifier with input and output impedances of 160 Ω; gain is 50–60 db over a bandwidth of 100 mc. Pulses with rise times as short as 4.2×10^{-9} sec are satisfactorily reproduced. Linearity is good up to 8v output.

621.375.232 2898

L.F. Compensation for Video Amplifiers: Part 2—Negative-Feedback Amplifiers—J. E. Flood. (*Wireless Eng.*, vol. 31, pp. 214–221; August, 1954.) The method outlined in Part 1 (2609 of August) is applied to negative-feedback amplifiers. If β is independent of frequency, the over-all response is optimum when the indicial response of the μ path is optimum and the order of compensation is the same. If β is made frequency dependent by including a capacitor in the feedback path, compensation can be obtained which is one order higher than that obtainable from the μ path alone. The method is applied to some simple circuits, including cases where a coupling or a decoupling circuit is outside the feedback loop. Experimental results confirming the theory are reported.

621.375.232.4.029.62:621.385.3:621.396.822 2899

Study of Noise Reduction by Feedback in Ultra-high Frequency Amplifiers—A. B. Glenn. (*Trans. I.R.E.*, No. PGBTR-6, pp. 82–119; April, 1954.) A detailed investigation has been made to determine theoretical and practical limits for the cancellation of tube noise in rf amplifiers. Measurements were made on a modified planar type of disk-seal triode designed to have a transit phase angle of about one radian at 120 mc. The conditions for complete noise cancellation are determined and expressions are derived for input and output admittance, available power gain (W) and noise figure (F). F can be reduced considerably by means of feedback but W is thereby reduced at an equivalent or greater rate, so that the practical application of this method is uneconomical with present tubes. The uncorrelated part of the induced grid noise is the most important factor in the increase in F at vhf.

621.375.3 2900

Single-Core Magnetic Amplifier Analysis—M. Frank, S. Rabotnick, and J. R. Walker. (*Elec. Eng.*, vol. 73, p. 336; April, 1954.) Digest of paper to be published in *Trans. AIEE, Communication and Electronics;* 1954.

621.375.4 2901

Dissected Amplifiers using Negative Resistance—W. Shockley and W. P. Mason. (*Jour. Appl. Phys.*, vol. 25, p. 677; May, 1954.) Possible circuit arrangements of semiconductor devices to give hf amplification are discussed, based on segregation of the amplifier properties of negative resistance and directionality. An arrangement of a hexagonal Ge plate with magnetic field perpendicular to its plane and with an appropriate network of resistances corresponds to a pentode amplifier.

621.375.4.029.4 2902

Transistor Application Fundamentals—R. F. Shea. (*Elec. Eng.*, vol. 73, pp. 360–365; April, 1954.) A comparison is made of tube and transistor characteristics in relation to af amplifier circuits. The design of a gramophone pickup amplifier using (*a*) the 62C7 twin-triode and (*b*) 2N43 Ge fused-junction transistors is described: circuit diagrams, characteristic curves and relevant formulas are given.

621.385.5 2903

Resonant Dielectric Amplifier Frequency Response—G. W. Penney, E. A. Sack, and E. R. Wingrove. (*Elec. Eng.*, vol. 73, p. 311; April, 1954. Digest of paper to be published in *Trans. AIEE, Communication and Electronics*; 1954.) Discussion of an amplifier of the type in which the capacitance of a nonlinear capacitor, excited to resonance by a carrier wave, is modulated by a signal voltage and an amplified output is obtained by subsequent demodulation. With a carrier-circuit Q value of 110, the carrier frequency must be of the order of 7 mc for the response to be flat up to signal frequencies of 10 kc.

621.376.23:538.221:538.652 2904

Detection of a Radio Wave by a Ferrite—J. C. Simon and G. Broussaud. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 2294–2296; June 14, 1954.] A magnetostrictive ferrite rod of 8-mm-square cross section and length 100 mm was exposed to a uhf field pulsed at 72 kc, the third harmonic of the frequency of mechanical resonance of the rod. The 72-kc signal was detected by means of a tuned coil located at a mechanical node. The detection process involves (a) rectification of the uhf oscillation as a result of the variation of ferrite permeability with applied field, and (b) conversion of the rectified signal into a magnetostrictive vibration. Detectors of this type may be useful at wavelengths of 3–10 cm.

621.396.6:534.321.9:534.213 2905

Characteristics of Ultrasonic Delay Lines using Quartz and Barium Titanate Ceramic Transducers—J. E. May, Jr. [*Jour. Acous. Soc. Amer.*, vol. 26, pp. 347–355; May, 1954. *Proc. NEC* (*Chicago*), vol. 9, pp. 264–277; 1953.] Equivalent-circuit-based analysis developed previously [2462 of 1952 (Fagen)] is extended to deal with metal delay media and with $BaTiO_3$ as well as quartz transducers. Calculations are made of the decrease in loss due to a decrease in the impedance of the delay medium and of the increase in bandwidth due to an increase in its impedance.

621.396.6.002.2 2906

Simplifying Fabrication of Printed Circuits—E. R. Gamson, A. Henesian, and R. W. Remington. (*Elec. Mfg.*, vol. 53, pp. 122–125; February, 1954.) Use of nitric acid in a stencil-etch process reduces production time greatly.

GENERAL PHYSICS

535.37:537.533 2907

Comparison between Luminescence and Exoelectron Emission—J. Kramer. (*Naturwissenschaften*, vol. 41, pp. 160–161; April, 1954.) The mechanism and significance of exoelectron emission (2301 of 1953) are investigated by studying the emission of light quanta by luminescence under the same conditions. Parallel effects in the two classes are tabulated and the close relation between the two is established.

535.42:538.566]+534.26 2908

Diffraction by Two Parallel Planes of Finite Length—W. E. Williams. (*Proc. Camb. Phil. Soc.*, vol. 50, part 2, pp. 309–318; April, 1954.) The diffracting system comprises a pair of parallel perfectly conducting planes of infinite width and finite length l. An incident em wave polarized perpendicular to the edges of the planes is considered; the theory thus applies equally to the diffraction of a plane sound wave. Two integral equations are derived; approximate solutions are obtained for the condition $l>\lambda$. An equation is derived giving the value of l for resonance of the system; this is solved approximately and a formula is given for the end correction, i.e. the difference between the resonant length and $n\lambda/2$.

535.42:538.566 2909

Diffraction Theory—C. J. Bouwkamp. (*Rep. Progr. Phys.*, vol. 17, pp. 35–100; 1954.) A critical review of progress in classical diffraction theory, including references to over 500 papers published since 1940. Both scalar and electromagnetic problems are discussed; the main emphasis is on diffraction by plane obstacles, especially by obstacles small compared with λ.

536.7:537.56 2910

Solution of Boltzmann's Equation for a Lorentz Gas: Application to Weakly Ionized Gases—M. Bayet, J. L. Delcroix and J. F. Denisse. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 2146–2148; May 31, 1954.]

536.8:537.226 2911

A Note on the Use of the Titanates as Thermoelectric Transducers—D. G. Frood. (*Canad. Jour. Phys.*, vol. 32, pp. 313–317; April, 1954.) Approximate calculations for a 75 per cent–25 per cent solid solution of $(Ba\text{-}Sr)TiO_3$ indicate that it would not be useful as a thermoelectric transducer. It seems possible that if materials could be produced with high permittivity and high temperature coefficient of permittivity but independent of field strength, a transducer with an efficiency comparable to that of a steam engine could be developed.

537.213+538.123]:517.947.42 2912

Green's Functions in Evaluating Fields—J. J. Smith. (*Elec. Eng.*, vol. 73, p. 323; April, 1954.) Digest of paper to be published in *Trans. AIEE, Communication and Electronics;* 1954. Analysis based on the derivation of vector potentials from scalar potentials (2285 of 1953) is extended to cases in which the field varies with time. An outline is given of the method used to calculate the field due to a current in a conductor. Applications include the calculation of fields in waveguides and resonant cavities.

537.312.8 2913

The Magnetoresistance Effect in Metals at High Frequencies—B. Donovan. (*Proc. Phys. Soc.*, vol. 67, pp. 305–314; April 1, 1954.) The calculation of the effect is based on a model in which the electrons occupy two overlapping partially filled bands of normal form. At low frequencies the relative change in surface resistance is independent of frequency and its variation with field strength is very similar to that observed in dc experiments. At high frequencies the effect ultimately drops to zero, the frequency region in which this occurs being determined by the relaxation time. For normal metals it lies in the far infrared. Experimental results for Bi suggest that the relaxation time in Bi is abnormally large.

537.52 2914

Current/Voltage Characteristics of a Self-Supporting Gas Discharge in a Transverse Magnetic Field—R. Haefer. (*Acta Phys. austriaca*, vol. 8, pp. 213–224; April, 1954.) Continuation of previous work (696 of March). Measurements were made using coaxial-cylinder electrodes with axial magnetic field, in argon at pressures from 10^{-1} to below 10^{-7} Torr. Falling or rising characteristics are obtained according as the pressure is greater or smaller than some thousandths of a Torr. At pressures above this value the discharge has the nature of a glow discharge, with positive space charge. At pressures below this value the characteristics are those of a self-supporting high-vacuum discharge, with negative space charge.

537.525 2915

Study of the Induced Electrodeless Discharge in Neon—F. Cabannes. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 1979–1981; May 17, 1954.] Induction-heating theory is used to study the conductivity in the induced discharge at a frequency of 1 mc.

537.525.029.64 2916

On the Radial Extent of Glow in Microwave Gas Discharge between Coaxial Cylinders—K. Mitani. [*Jour. Phys. Soc.* (*Japan*), vol. 9, pp. 299–300; March/April, 1954.] The glow appears in the region of maximum electric field strength close to the central wire, over a radius ~1–2 mm. The luminosity depends on the net production rate of electrons per unit volume. This rate is calculated and is shown graphically as a function of radius.

537.527.2 2917

Further Research on the Discharge between Positive Point and Plane in Compressed Air or CO_2—A. Boulloud. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 1981–1983; May 17, 1954.] Continuation of work described previously [3265 of 1953 (Toitot and Boulloud)].

537.533.8:546.561 2918

Investigation of the Energy Distribution of Secondary Electrons from Single Crystals of Copper Covered with Single-Crystal Layers of Copper Oxide—N. B. Gornyi. (*Zh. eksp. teor. Fiz.*, vol. 26, pp. 327–336; March, 1954.) The secondary electron energy was measured by the retarding-field method. The energy-distribution curve is steeper towards the low-energy side. In the case of polycrystalline Cu covered with polycrystalline Cu_2O one well defined maximum was observed; in the single-crystal case subsidiary maxima were also found, but the evidence for these is not conclusive. The results are shown graphically.

537.533.8:621.387 2919

A Reversible Secondary-Emission Fatigue Effect observed in Glow Cathodes comprising a Semiconducting Layer—W. Kluge and A. Schulz. (*Z. Phys.*, vol. 137, pp. 392–400; April 14, 1954.) If the usual metallic cathode in a glow-discharge tube is replaced by a composite cathode having a semiconducting layer, e.g. an Ag-Cs_2O, (Ag), (Cs)-Cs cathode, the self-maintained discharge current decreases with time, when the current density does not exceed a particular value dependent on cathode geometry, and when the cell is kept in the dark. Recovery takes place slowly, but is speeded up if the cell is subjected to infrared radiation. The effect is ascribed to reversible fatigue of the emission centers for secondary electrons, and the analogous effect observed with photocathodes suggests that both photons and secondary electrons originate in the same emission centers.

538.11 2920

Antiferromagnetism—A. B. Lidiard. (*Rep. Progr. Phys.*, vol. 17, pp. 201–244; 1954.) A survey paper giving particular attention to Van Vleck's theory and including a discussion of the superexchange mechanism. Papers on experimental investigations are listed by compound and by property studied. Nearly 200 references.

538.242 2921

Fisher's Gyromagnetic Effect observed in a Body Placed in a Waveguide—K. M. Polivanov. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 95, pp. 969–970; April 11, 1954. In Russian.] The component of magnetization produced in a ferrite perpendicular to the plane of a rotating magnetizing field may be detected in an ellipsoidal specimen placed in a waveguide excited in the H_{01} mode. The theory and experiments which gave negative results were discussed by Fisher (*Proc. Roy. Soc. A*, vol. 109, pp. 7–27; September 1, 1925.) The effect can be used in determining magnetic fields at very high frequencies.

538.3 2922

Interpretation of the Ampère Experiments—P. Moon and D. E. Spencer. (*Jour. Frank. Inst.*, vol. 257, pp. 203–220; March, 1954.) Ampère's equation for the force between current elements is unique, assuming the correctness of his five postulates. The general equations are here derived that apply if either the third or the fifth postulate is omitted.

538.3 2923

The Coulomb Force and the Ampère Force—P. Moon and D. E. Spencer. (*Jour. Frank. Inst.*, vol. 257, pp. 305–315; April, 1954.) A study of the modification of Coulomb's equation that will satisfy Ampère's experimental results.

538.3 2924

A New Electrodynamics—P. Moon and D. E. Spencer. (*Jour. Frank. Inst.*, vol. 257, pp. 369–382; May, 1954.) A new formulation, alternative to that of Maxwell's equations, is developed from the theories of Weber and of Ritz. The fundamental equation expresses the force on a charged particle caused by another charged particle. For preliminary work *see* 2922 and 2923 above.

538.56:537.562 2925

Electromagnetic Properties of Plasmas in a Magnetic Field—M. Bayet. (*Jour. Phys. Radium*, vol. 15, pp. 258–263; April, 1954.) Continuation of work reported in 1659 of 1953, assuming a constant and uniform magnetic field. Results obtained are discussed in relation to those of other workers, in particular, the matrix for the complex conductivity of an ionized gas, and the dispersion equation. Previous results apply only for values of gyromagnetic frequency small compared with $(f+i\omega)$, where f is the electron collision frequency and ω the pulsatance of the electric field.

538.566:534.21-14 2926

Magneto-Hydrodynamic Waves in Liquid Sodium—B. Lehnert. (*Phys. Rev.*, vol. 94, pp. 815–824; May 15, 1954.) Torsional waves in liquid Na have been generated in a cylindrical vessel with the axis parallel to a homogeneous magnetic field, and resonance phenomena have been investigated.

538.566:538.614 2927

Propagation of Electromagnetic Waves in a Gyrotropic Medium—M. A. Gintsburg. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 95, pp. 753–756; April 1, 1954. In Russian.] Faraday's law of rotation of the plane of polarization is extended to take account of the reflection and refraction at the boundaries of the medium. The relation between path length and the angle of rotation is derived and results for a layer of $MnO \cdot ZnO \cdot 2Fe_2O_3$ are shown graphically for a frequency of 2×10^{10} cps.

538.639 2928

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GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.165 2929

Theory of Origin of Cosmic Rays—E. Teller. (*Rep. Progr. Phys.*, vol. 17, pp. 154–172; 1954.) "It is shown that a random acceleration process, proposed by Fermi, and a physical location of the accelerating fields in radio stars, proposed by Unsold, account satisfactorily for the main features of cosmic radiation."

523.53:621.396.96 2930

Radar Determination of the Heights and Radiants of Meteor Showers—D. W. R. McKinley. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 2224–2225; June 9, 1954.] The technique described by Volmer (713 of March) was used for observations of the Geminids and δ-Aquarids. The results are discussed briefly.

523.755 2931

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523.854:621.396.822.029.63 2932

Probable Observation of the Galactic Nucleus at 400 Mc/s—R. X. McGee and J. G. Bolton. [*Nature* (*London*), vol. 173, pp. 985–987; May 22, 1954.] Brief report of a survey of the central section of the Milky Way, using an antenna system with a beamwidth of 2 degrees. In the zone covered from −17 degrees to −49 degrees in declination and from 15 hours to 20 hours in right ascension, a strong concentration of brightness was found near a line of galactic latitude −1 degrees with a maximum centered at 327.9 degrees longitude.

551.508:621.398 2933

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551.510.5:621.396.11 2934

N.B.S. Research in Radio Propagation—(*See* 3018.)

551.510.535 2935

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551.510.535 2936

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551.510.535:523.78 2940

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551.510.535:550.38 2941

Polar Magnetic Storms and Geomagnetic Bays: Appendix 1—A Theory of S_D Field—N. Fulcushima and T. Oguti. [*Rep. Ionosphere Res. (Japan)*, vol. 7, pp. 137–146; December, 1953.) An S_D current system deduced from geomagnetic observations is compared with a system calculated on the dynamo theory, taking anisotropic conductivity of the ionosphere into account. Agreement is fairly good. Conductivity considerations suggest that the S_D current system is located at a height below 150 km.

551.510.535:550.384 2942

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551.510.535:551.55 2944

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551.594 2947

Atmospheric Electricity—J. A. Chalmers. (*Rep. Progr. Phys.*, vol. 17, pp. 101–134; 1954.) A review paper. Aspects of the subject discussed include some in which significant advances have been made since publication of the author's book (98 of 1950) and some on which work was done in Germany during the war. About 100 references.

551.594.6:538.566.029.42 2948

Observation of Atmospherics at Very Low Frequencies—W. O. Schumann and H. König. (*Naturwissenschaften*, vol. 41, pp. 183–184; April, 1954.) Results of observations, at Munich and several other places, at frequencies in the range 1–25 cps are correlated with meteorological phenomena. Two main types of waveform were distinguished, one comprising wave trains of duration 0.3–30 sec at a frequency about 9 cps, the other of a generally nonsinusoidal character at frequencies between 3 and 6 cps. The former is associated with the presence of a temperature inversion, the latter with low nimbus clouds. The latter type occurred most frequently one to two days prior to the onset of increased temperatures. Neither type of atmospheric could be observed following the onset of Föhn-type weather.

LOCATION AND AIDS TO NAVIGATION

621.396.932.25:621.396.96 2949

A Racon Beacon for Reception by Civil Marine Radars—C. Randall-Cook. (*Jour. Inst. Nav.*, vol. 7, pp. 181–186; April, 1954. Discussion, pp. 190–194.) The beacon is triggered by any radar signal in the 3-cm band at horizon range. On the ship's display, the position of the beacon is identified by a bright radial line starting from the beacon range. Beacon bearing is thus indicated without obscuring the display at closer ranges. No modification to existing marine equipment is required. Beacon equipment comprises a transmitter with frequency sweep over the range 9.320–9.500 kmc, a receiver for the same range, and transmitting and receiving antennas having omnidirectional azimuth coverage and 30-degree vertical coverage to allow for roll and pitch when fitted to a light vessel.

621.396.96 2950

False Alarm Time in Pulse Radar—R. Hollis. (Proc. I.R.E., vol. 42, p. 1189; July, 1954.) Two different definitions of false-alarm time are examined. Though the one definition yields a time 45 per cent longer than the other, the probabilities of one false alarm occurring during these times differ by only about 6 per cent.

621.396.96:551.578.1 2951

Effect of Attenuation on the Choice of Wavelength for Weather Detection by Radar—W. Hitschfeld and J. S. Marshall. (Proc. I.R.E., vol. 42, pp. 1165–1168; July, 1954.) Decreasing the wavelength increases the radar range in clear air but also increases the attenuation due to rain. For the certain detection of operationally significant storms, wavelengths below about 5 cm are unsuitable. See also 2404 of August (Hitschfeld and Bordan).

621.396.96:621.396.932.25 2952

A Beacon for Ship Identification—A. L. P. Milwright. (*Jour. Inst. Nav.*, vol. 7, pp. 187–190; April, 1954. Discussion, pp. 190–194.) Ships entering harbor with the assistance of surveillance radar must be identified on the radar display. The ship station for this purpose consists of vhf transponder equipment weighing 23 pounds, taken aboard by the pilot. The unit has directive antennas and built-in power supplies with a life of 40 minutes continuous operation. The type of signal chosen is a short radial line passing through the ship range, with a gap in the line at that range. The gap is produced by permitting the beacon klystron to oscillate continuously and using the received signal to stop the oscillation for a period of 12 μs. This beacon is not considered satisfactory for ship-to-ship identification.

621.396.963 2953

The Reduction of Sea and Rain Clutter in Marine Radars—J. Croney. (*Jour. Inst. Nav.*, vol. 7, pp. 175–180; April, 1954. Discussion, pp. 190–194.) For low-definition S-band radar receivers, sea or rain clutter has an amplitude probability distribution of approximately Rayleigh type. When clutter or noise signals are applied to a logarithmic receiver, the rms amplitude, at the output, of clutter or noise fluctuations about a mean value is constant and independent of that mean value. It is then merely necessary to eliminate the mean value (e.g. by a differentiating circuit) in order to reduce all clutter intensity to the level of the inherent receiver noise. The slope of the receiver logarithmic characteristic must be constant down to about 20 db below rms noise level. Results on high-definition X-band radar are less good, since the amplitude probability distribution law is not the same.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.7 2954

A Demountable All Metal Hot-Cathode Vacuum Ionisation Gauge—P. K. Dutt. (*Indian Jour. Phys.*, vol. 28, pp. 1–8; January, 1954.) Details are given of the construction and performance of a gauge whose filament can be changed without affecting the operating characteristics.

531.788.7 2955

An Ionization Manometer and Control Unit for Extremely Low Pressures—P. A. Red-

537.533.8:621.387 2919
A Reversible Secondary-Emission Fatigue Effect observed in Glow Cathodes comprising a Semiconducting Layer—W. Kluge and A. Schulz. (*Z. Phys.*, vol. 137, pp. 392–400; April 14, 1954.) If the usual metallic cathode in a glow-discharge tube is replaced by a composite cathode having a semiconducting layer, e.g. an Ag-Cs_2O, (Ag), (Cs)-Cs cathode, the self-maintained discharge current decreases with time, when the current density does not exceed a particular value dependent on cathode geometry, and when the cell is kept in the dark. Recovery takes place slowly, but is speeded up if the cell is subjected to infrared radiation. The effect is ascribed to reversible fatigue of the emission centers for secondary electrons, and the analogous effect observed with photocathodes suggests that both photons and secondary electrons originate in the same emission centers.

538.11 2920
Antiferromagnetism—A. B. Lidiard. (*Rep. Progr. Phys.*, vol. 17, pp. 201–244; 1954.) A survey paper giving particular attention to Van Vleck's theory and including a discussion of the superexchange mechanism. Papers on experimental investigations are listed by compound and by property studied. Nearly 200 references.

538.242 2921
Fisher's Gyromagnetic Effect observed in a Body Placed in a Waveguide—K. M. Polivanov. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 95, pp. 969–970; April 11, 1954. In Russian.] The component of magnetization produced in a ferrite perpendicular to the plane of a rotating magnetizing field may be detected in an ellipsoidal specimen placed in a waveguide excited in the H_{01} mode. The theory and experiments which gave negative results were discussed by Fisher (*Proc. Roy. Soc. A*, vol. 109, pp. 7–27; September 1, 1925.) The effect can be used in determining magnetic fields at very high frequencies.

538.3 2922
Interpretation of the Ampère Experiments—P. Moon and D. E. Spencer. (*Jour. Frank. Inst.*, vol. 257, pp. 203–220; March, 1954.) Ampère's equation for the force between current elements is unique, assuming the correctness of his five postulates. The general equations are here derived that apply if either the third or the fifth postulate is omitted.

538.3 2923
The Coulomb Force and the Ampère Force—P. Moon and D. E. Spencer. (*Jour. Frank. Inst.*, vol. 257, pp. 305–315; April, 1954.) A study of the modification of Coulomb's equation that will satisfy Ampère's experimental results.

538.3 2924
A New Electrodynamics—P. Moon and D. E. Spencer. (*Jour. Frank. Inst.*, vol. 257, pp. 369–382; May, 1954.) A new formulation, alternative to that of Maxwell's equations, is developed from the theories of Weber and of Ritz. The fundamental equation expresses the force on a charged particle caused by another charged particle. For preliminary work *see* 2922 and 2923 above.

538.56:537.562 2925
Electromagnetic Properties of Plasmas in a Magnetic Field—M. Bayet. (*Jour. Phys. Radium*, vol. 15, pp. 258–263; April, 1954.) Continuation of work reported in 1659 of 1953, assuming a constant and uniform magnetic field. Results obtained are discussed in relation to those of other workers, in particular, the matrix for the complex conductivity of an ionized gas, and the dispersion equation. Previous results apply only for values of gyromagnetic frequency small compared with $(f+i\omega)$, where f is the electron collision frequency and ω the pulsatance of the electric field.

538.566:534.21-14 2926
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Atmospheric Electricity—J. A. Chalmers. (*Rep. Progr. Phys.*, vol. 17, pp. 101–134; 1954.) A review paper. Aspects of the subject discussed include some in which significant advances have been made since publication of the author's book (98 of 1950) and some on which work was done in Germany during the war. About 100 references.

551.594.6:538.566.029.42 2948

Observation of Atmospherics at Very Low Frequencies—W. O. Schumann and H. König. (*Naturwissenschaften*, vol. 41, pp. 183–184; April, 1954.) Results of observations, at Munich and several other places, at frequencies in the range 1–25 cps are correlated with meteorological phenomena. Two main types of waveform were distinguished, one comprising wave trains of duration 0.3–30 sec at a frequency about 9 cps, the other of a generally nonsinusoidal character at frequencies between 3 and 6 cps. The former is associated with the presence of a temperature inversion, the latter with low nimbus clouds. The latter type occurred most frequently one to two days prior to the onset of increased temperatures. Neither type of atmospheric could be observed following the onset of Föhn-type weather.

LOCATION AND AIDS TO NAVIGATION

621.396.932.25:621.396.96 2949

A Racon Beacon for Reception by Civil Marine Radars—C. Randall-Cook. (*Jour. Inst. Nav.*, vol. 7, pp. 181–186; April, 1954. Discussion, pp. 190–194.) The beacon is triggered by any radar signal in the 3-cm band at horizon range. On the ship's display, the position of the beacon is identified by a bright radial line starting from the beacon range. Beacon bearing is thus indicated without obscuring the display at closer ranges. No modification to existing marine equipment is required. Beacon equipment comprises a transmitter with frequency sweep over the range 9.320–9.500 kmc, a receiver for the same range, and transmitting and receiving antennas having omnidirectional azimuth coverage and 30-degree vertical coverage to allow for roll and pitch when fitted to a light vessel.

621.396.96 2950

False Alarm Time in Pulse Radar—R. Hollis. (Proc. I.R.E., vol. 42, p. 1189; July, 1954.) Two different definitions of false-alarm time are examined. Though the one definition yields a time 45 per cent longer than the other, the probabilities of one false alarm occurring during these times differ by only about 6 per cent.

621.396.96:551.578.1 2951

Effect of Attenuation on the Choice of Wavelength for Weather Detection by Radar—W. Hitschfeld and J. S. Marshall. (Proc. I.R.E., vol. 42, pp. 1165–1168; July, 1954.) Decreasing the wavelength increases the radar range in clear air but also increases the attenuation due to rain. For the certain detection of operationally significant storms, wavelengths below about 5 cm are unsuitable. See also 2404 of August (Hitschfeld and Bordan).

621.396.96:621.396.932.25 2952

A Beacon for Ship Identification—A. L. P. Milwright. (*Jour. Inst. Nav.*, vol. 7, pp. 187–190; April, 1954. Discussion, pp. 190–194.) Ships entering harbor with the assistance of surveillance radar must be identified on the radar display. The ship station for this purpose consists of vhf transponder equipment weighing 23 pounds, taken aboard by the pilot. The unit has directive antennas and built-in power supplies with a life of 40 minutes continuous operation. The type of signal chosen is a short radial line passing through the ship range, with a gap in the line at that range. The gap is produced by permitting the beacon klystron to oscillate continuously and using the received signal to stop the oscillation for a period of 12 μs. This beacon is not considered satisfactory for ship-to-ship identification.

621.396.963 2953

The Reduction of Sea and Rain Clutter in Marine Radars—J. Croney. (*Jour. Inst. Nav.*, vol. 7, pp. 175–180; April, 1954. Discussion, pp. 190–194.) For low-definition S-band radar receivers, sea or rain clutter has an amplitude probability distribution of approximately Rayleigh type. When clutter or noise signals are applied to a logarithmic receiver, the rms amplitude, at the output, of clutter or noise fluctuations about a mean value is constant and independent of that mean value. It is then merely necessary to eliminate the mean value (e.g. by a differentiating circuit) in order to reduce all clutter intensity to the level of the inherent receiver noise. The slope of the receiver logarithmic characteristic must be constant down to about 20 db below rms noise level. Results on high-definition X-band radar are less good, since the amplitude probability distribution law is not the same.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.7 2954

A Demountable All Metal Hot-Cathode Vacuum Ionisation Gauge—P. K. Dutt. (*Indian Jour. Phys.*, vol. 28, pp. 1–8; January, 1954.) Details are given of the construction and performance of a gauge whose filament can be changed without affecting the operating characteristics.

531.788.7 2955

An Ionization Manometer and Control Unit for Extremely Low Pressures—P. A. Red-

head and L. R. McNarry. (*Canad. Jour. Phys.*, vol. 32, pp. 267–274; April, 1954.) A gauge of the Bayard-Alpert type is described, for the measurement of pressures as low as 10^{-10} mm Hg; methods for increasing the sensitivity are discussed.

535.37 2956

Intensity of Luminescence of Powdered Luminophors—A. P. Ivanov. (*Zh. eksp. teor. Fiz.*, vol. 26, pp. 275–280; March, 1954.) Theoretical calculation of the dependence of the intensity of luminescence on the degree of dispersion, the binding medium and the thickness of luminescent layer.

535.37 2957

Effect of Neutron Bombardment on a Zinc Sulfide Phosphor—A. W. Smith and J. Turkevich. (*Phys. Rev.*, vol. 94, pp. 857–865; May 15, 1954.) Luminescence efficiency after bombardment decreased for all conditions of temperature, the effect being more pronounced the longer the period of neutron irradiation. The emission spectrum was practically unchanged. Bombardment changes neither the permittivity nor its dependence on excitation light intensity.

535.37:061.3 2958

Luminescence or Inorganic Phosphors—G. F. J. Garlick. [*Nature* (*London*), vol. 173, pp. 1077–1078; June 5, 1954.] Brief account of the proceedings of a conference held at Cambridge in April, 1954.

535.37:537.228 2959

New Effect of Sensitization of Luminescent Substances to X-Rays by Electric Fields—M. Destriau. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 2298–2300; June 14, 1954.] In contrast to the quenching effect described by G. Destriau (*Jour. Phys. Radium*, vol. 4, p. 32; 1943) certain mixtures of ZnS and CdS when excited by X-rays exhibit an increase of luminescence on application of an electric field. The same mixtures exhibit the usual quenching effect when excited by ultraviolet rays.

535.37:537.529:546.482.21 2960

Electric Breakdown and Light Emission in CdS Single Crystals—G. Diemer. (*Philips Res. Rep.*, vol. 9, pp. 109–121; April, 1954.)

535.376 2961

New Manganese-Activated Fluoride Phosphors—A. L. Smith. (*Jour. Electrochem. Soc.*, vol. 101, pp. 189–194; April, 1954.) "The following manganese-activated fluorides were found to exhibit efficient cathodoluminescence: $NaZnF_3$, $KZnF_3$, K_2ZnF_4, $NaMgF_3$, $KMgF_3$, K_2MgF_4, CaF_2, $KCaF_3$, $KCdF_3$, and the solid solutions formed between CaF_2 and AlF_3."

535.376:546.472.21 2962

Electroluminescent ZnS Phosphors—P. Zalm, G. Diemer, and H. A. Klasens. (*Philips Res. Rep.*, vol. 9, pp. 81–108; April, 1954.) An account is given of the preparation of a ZnS phosphor with a copper-rich layer on the surface of the grains. Electrical and optical measurements made on such phosphors are described; voltage/current and voltage/light-output oscillograms for suspended and binderless settled layers are shown. The maximum efficiency is 1.5 ± 0.2 lm/W, the brightness at 600v, 3 kc is $\sim$1000 lm/m^2. The hypothesis of a barrier action at the grain surface was confirmed. Results of observations lead to the conclusion that electroluminescence is similar to cathodoluminescence with "slow" electrons. The low efficiency of electroluminescent phosphors is compared with that of normal phosphors under cathode-ray excitation at low voltage and high current density. An equivalent circuit for a suspended electroluminescent layer is discussed.

537.226.33+539.389.3 2963

Dielectric Hysteresis and Elastic After-Effect—J. Granier and P. Caillon. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 2232–2233; June 9, 1954.] The relations between the two effects were investigated by measurements on polyvinyl chloride. Delay intervals studied ranged from 5 to 2,000 seconds, and temperatures covered the whole range of states of viscoelasticity of the substance. The results indicate different and independent origins for the two effects over the time intervals studied.

537.227 2964

Theory of Ferroelectrics—A. F. Devonshire. (*Advances Phys.*, vol. 3, pp. 85–130; April, 1954.) Ferroelectric substances are defined and classified, their physical constants, their properties near the transition temperature, and the relation of their macroscopic properties to atomic structure are examined.

537.227:547.476.3 2965

Nature of Ferroelectricity in Rochelle Salt—K. N. Karmen. (*Zh. eksp. teor. Fiz.*, vol. 26, pp. 370–376; March, 1954.) Results of an experimental investigation of the charging and discharging of a capacitor with the salt as dielectric indicate that the time lags observed are not due to an intermediate layer between the electrodes and the dielectric, nor can they be explained in terms of the usual theory of ferroelectricity. The results of measurements at temperatures between about -18 degrees C. and $+20$ degrees C. and at field strengths up to 280v/cm are shown graphically.

537.311.3 2966

Electrical Conduction in Halide-Contaminated Ice—E. J. Workman, F. K. Truby, and W. Drost-Hanson. (*Phys. Rev.*, vol. 94, p. 1073; May 15, 1954.) Halide-contaminated ice, frozen in a prescribed way, has the following characteristics: (a) it acts as a rectifier, (b) it has an extremely high permittivity for fields applied in the inverse direction, and (c) the conductive and dielectric properties depend in a complicated way upon temperature, bias, and frequency. Experiments show that the conduction process is ionic, involving a proton transfer mechanism.

537.311.33:535.215 2967

Kinetics of the Formation and Relaxation of Non-equilibrium Current Carriers on Illumination of a Semiconductor—E. I. Adirovich and V. G. Kolomilova. (*Zh. eksp. teor. Fiz.*, vol. 26, pp. 281–292; March, 1954.) A theoretical investigation for a semiconductor with bipolar conductivity. Solutions are found for a uniform layer, as solution of a linearized problem in a general form and in asymptotic forms for given boundary conditions. The stationary carrier-distribution is also determined. The conditions under which the linearized equations are applicable and their connection with the requirement of quasi-neutrality are discussed.

537.311.33:536.21 2968

On the Thermal and Electrical Conductivity of Semiconductors—H. J. Goldsmid. (*Proc. Phys. Soc.*, vol. 67, pp. 360–363; April 1, 1954. Correction, *ibid.*, No. 415 B; p. 599.) The relation between electron mobility μ and thermal conductivity λ is investigated for high-purity semiconductors, for which it may be assumed that scattering is entirely due to lattice vibrations and that thermal current is carried solely by elastic waves. Theory indicates a rapid rise of μ/λ with atomic weight; this is borne out by observations.

537.311.33:539.231:546.48–31 2969

Irreversible Changes in Electron Mobility in Thin CdO Films—F. Lappe. (*Z. Phys.*, vol. 137, pp. 380–382; April 14, 1954.) For films sputtered in air or in argon-oxygen mixtures, and heated in air, the variation of conductivity measured over the temperature range 20 degrees to 700 degrees C. is irreversible and the conductivity exhibits a minimum around 270 degrees C. Hall-effect measurements show that this is due to a minimum in electron mobility. This minimum is not observed for films heated in vacuum.

537.311.33:539.231:546.48–31 2970

Optical Absorption and Electrical Properties of CdO Films with Many Lattice Imperfections—H. Dunstädter. (*Z. Phys.*, vol. 137, pp. 383–385; April 14, 1954.) In addition to the electrical measurements reported in 2969 above, optical absorption measurements were made on the same films. Results show that the extinction coefficient decreases as the electron concentration increases.

537.311.33:546.23 2971

A Macrostructure of Amorphous Selenium—J. Weber. (*Z. Naturf.*, vol. 8a, pp. 565–566A; September, 1953.) The conditions are discussed under which evaporated layers of amorphous Se exhibit simulated crystal structure. Photographs at various magnifications illustrate the effect.

537.311.33:546.23 2972

Electrical Resistivity of Polycrystalline Selenium at Higher Fields—M. Kuwagata, E. Otsuka, and M. Tomura. [*Jour. Phys. Soc.* (*Japan*), vol. 9, pp. 300–301; March/April, 1954.] Measurements were made on a specimen annealed at $\sim$180 degrees C for 10 minutes, using electric fields of up to 22 kv/cm applied in 7-μs pulses at a repetition rate of 60/seconds. The resistivity/field-strength characteristic obtained fits the Schottky plot at temperatures of 40 degrees, 67 degrees and 100 degrees C. The variation of resistance with frequency up to 20 mc is shown graphically. For specimens containing 0.05 per cent of iodine, the resistivities are lower by a factor of 10^2–10^3 than those for a pure sample.

537.311.33:546.28:538.529 2973

Avalanche Breakdown in Silicon—K. G. McKay. (*Phys. Rev.*, vol. 94, pp. 877–884; May 15, 1954.) Theory is based on the assumption of approximately equal ionization rates for electrons and holes. The problem of determining ionization rates from data obtained in inhomogeneous fields is solved exactly for junctions in which the impurity concentration varies (a) abruptly, and (b) gradually, from p to n type. Ionization rates thus calculated from experimental data on breakdown voltage and on pre-breakdown multiplication are in good agreement in both cases. The temperature coefficient of the ionization rate exhibits a similar internal consistency. Breakdown in Si may therefore be regarded as a direct result of multiplication by collision. Detailed observations of the pulse-type noise associated with the onset of breakdown are reported.

537.311.33:546.289 2974

Interferometric Observation of Mosaic Structure on the (111) Face of a Single Crystal of Germanium—A. R. Verma. (*Proc. Phys. Soc.*, vol. 67, pp. 359–360; April 1, 1954.)

537.311.33:546.289 2975

Hall Coefficient in Germanium—T. C. Harman, R. K. Willardson, and A. C. Beer. (*Phys. Rev.*, vol. 94, pp. 1065–1067; May 15, 1954.) By carrying out measurements in strong magnetic fields and using the equations valid in the limiting cases of such fields, mobility values were obtained for p-type Ge in good agreement with drift-mobility measurements. The extrinsic Hall coefficient determined in strong magnetic fields remains relatively constant with decreasing temperature.

537.311.33:546.289 2976

High-Field Effects in Germanium—E. M. Conwell. (*Phys. Rev.*, vol. 94, p. 1068; May 15, 1954.) The experimental results of Ryder (3326 of 1953), interpreted by him as due to an increase in mobility, were alternatively explained by Sclar et al. (*Phys. Rev.*, vol. 92, p. 858; 1953), as due to impact ionization of neutral

donors by hot electrons. An experimental check is suggested.

537.311.33:546.289 2977
Volume and Surface Recombination Rates for Injected Carriers in Germanium—J. P. McKelvey and R. L. Longini. (*Jour. Appl. Phys.*, vol. 25, pp. 634–641; May, 1954.) Recombination is studied, using experimental technique described by Navon et al. (740 of 1953). Diffusion theory is developed, taking account of reflection of carriers at the surfaces of the sample. A simple relation is derived between the reflection coefficient and the surface recombination velocity. Values of bulk lifetime and surface recombination velocity are determined for *p*-type and *n*-type samples, for various surface treatments.

537.311.33:546.289 2978
Temperature Dependence of the Surface Recombination Velocity in Germanium—Y. Kanai. [*Jour. Phys. Soc.* (*Japan*), vol. 9, pp. 292–293; March/April, 1954.] The surface recombination velocity determined from measurements of the Suhl effect in *n*-type Ge is given as a function of temperature in the range from about −30 degrees C. to +60 degrees C. A tentative value of 0.35 eV is given for the energy level of the surface traps below the conduction band.

537.311.33:546.289:621.314.632 2979
Blocking Contacts between Germanium and Evaporated Metal Films—H. U. Harten. (*Naturwissenschaften*, vol. 41, p. 162; April, 1954.) Measurements were made of the differential capacitance of rectifying contacts prepared by evaporating Au or Cu on the previously etched Ge disks. The variation with applied blocking voltage agrees qualitatively but not quantitatively with that expected from theory. The rate of variation appears to depend not only on the impurity concentration but also, in a manner not yet understood, on the state of the Ge surface.

537.311:546.46–31 2980
Some Optical Properties of MgO in the Vacuum Ultraviolet—P. D. Johnson. (*Phys. Rev.*, vol. 94, pp. 845–846; May 15, 1954.) "The optical absorption and the excitation spectrum for luminescence of synthetic periclase crystals have been determined in the vacuum ultraviolet. From these measurements it is deduced that the width of the forbidden energy band of MgO is approximately 7.4 eV."

535.311.33:546.682.86 2981
The Electrical Properties of Indium Antimonide at Low Temperatures—J. Hatton and B. V. Rollin. (*Proc. Phys. Soc.*, vol. 67, pp. 385–386; April 1, 1954.) Measurements were made on polycrystalline specimens over the temperature range 300 degrees to 1 degree K. Hall coefficient and magnetoresistance are very small at the lowest temperatures.

537.311.33:[546.817.231+546.817.241 2982
Electrical Properties of Single Crystals and Thin Films of PbSe and PbTe—S. J. Silverman and H. Levinstein. (*Phys. Rev.*, vol. 94, pp. 871–876; May 15, 1954.) Thin films were formed by evaporating single crystals without exposure to air. Measurements were made in the temperature range 300 degrees to 77 degrees K. All films were *n* type, whether the original crystal was *n* or *p* type. Carrier mobilities in films were markedly affected by substrate temperature during film formation, subsequent, annealing, and limited oxygenation; room-temperature mobilities in annealed films were about $\frac{1}{8}$ of the bulk values.

538.221 2983
Surface Magnetic Anisotropy and Superlattices formed by Orientation—L. Néel. (*Jour. Phys. Radium*, vol. 15, pp. 225–239; April, 1954.) Theory showing the relation between these two effects is applied, in particular, to the interpretation of the phenomena observed by Rathenau and Snoek (3428 of 1948).

538.221 2984
Initial Complex Magnetic Permeability of Iron from 0 to 7000 Mc/s—J. Benoit and E. Naschke. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 2292–2294; June 14, 1954.] Permeability measurements were made on Armco wire treated in various ways. For a specimen treated to produce large domains, magnetic dispersion starts at a frequency of about 5 mc, i.e. well below the frequency at which electron spin rotation is effective. This enables the experimental results to be used as a check against Néel's theory of the movements of domain walls.

538.221 2985
On the *g*-Factor and H_i of Nickel Ferrite-Aluminate—Y. Torizuka, Y. Kojima, and T. Fujimura. [*Jour. Phys. Soc.* (*Japan*) vol. 9, pp. 298–299; March/April, 1954.] A note giving the results of experimental determinations of the *g* factor at 19.23 kmc and 28.85 kmc, taking account of the internal magnetic field.

538.221 2986
Cation Distributions in Mixed Ferrites—J. S. Smart. (*Phys. Rev.*, vol. 94, pp. 847–850; May 15, 1954.) Expressions are derived for the equilibrium distribution of cations in mixed ferrites having three different kinds of metallic ions, in terms of magnetic moment and *g* factor. For nickel ferrite aluminates the results agree reasonably well with X-ray diffraction data.

538.614:[621.318.124+621.318.134 2987
Faraday Effect in Magnetic Materials with Travelling and Standing Waves—H. G. Beljors. (*Philips Res. Rep.*, vol. 9, pp. 131–139; April, 1954.) The measurement of Faraday rotation at 9.4 and 24 kmc in ferroxcube and in ferroxdure cylinders in circular waveguides is described. The effect was also demonstrated by the splitting up of the resonance curve of a cavity resonator at one end of which was placed a thin disk of ferroxcube IV subjected to a variable magnetic field.

538.652 2988
The Approach to Saturation Magnetostriction—E. W. Lee. (*Proc. Phys. Soc.*, vol. 67, pp. 381–383; April 1, 1954.) Magnetostriction of a polycrystalline specimen is given in terms of the external field, H, by $(A+BH^{-1}+CH^{-2}+\ldots)$. The conditions under which the term in H^{-2} is of real importance are discussed.

546.311.12 2989
Color Centers in Alkali Halide Crystals: Part 2—F. Seitz. (*Rev. Mod. Phys.*, vol. 26, pp. 7–94; January, 1954.) A survey of progress since 1946

548.55:546.289 2990
Cross Section of Pulled Crystals—R. G. Pohl. (*Jour. Appl. Phys.*, vol. 25, pp. 668–669; May, 1954.) From considerations of mechanical and thermal equilibrium at the liquid/solid interface, an expression is derived for the radius of a crystal grown by the Czochralski technique. The values obtained are compared with experimental results for Ge crystals.

621.318.23:621.385.029.6 2991
Calculation of Permanent Magnets of "Tubular" Form—M. de Bennetot. (*Ann. Radioelect.*, vol. 9, pp. 193–216; April, 1954.) The design of magnets for focusing the beams of uhf tubes is considered. A procedure based on the equivalence between the magnet and a system of fictitious magnetic masses is studied. From the geometry of the system and the demagnetization characteristic of the material, the field along the axis of revolution is determined by means of a relaxation method. Calculated values are in good agreement with values found experimentally.

621.319.4:621.793 2992
The Zinc Coating of Paper for Capacitors by Vacuum Evaporation—L. Holland and K. Hacking. (*Elec. Eng.*, vol. 26, pp. 296–301; July, 1954.) An account is given of the development of equipment and technique for producing the low gas pressure required for metallizing volatile materials such as capacitor tissue. *Zn* can be evaporated at higher rates and at higher gas pressures than *Al*, but when using *Zn* it is necessary to pre-coat the paper with nucleating film, e.g. *Ag* monolayer. To avoid oxidation of the *Zn* coating, the paper must be thoroughly pre-dried.

MATHEMATICS

517 2993
New Formulas for Facilitating Osculatory Interpolation—H. E. Salzer. (*Jour. Res. Nat. Bur. Stand.*, vol. 52, pp. 211–216; April, 1954.) Hermite's *n*-point osculatory interpolation formula for equally spaced arguments, employing the function and its derivative, is expressed in a form which reduces the amount of labor required to calculate the coefficients.

517.564.3 2994
Asymptotic Expansions of Bessel Functions in the Transition Regions—M. Nagase. [*Jour. Phys. Soc.* (*Japan*), vol. 9, pp. 296–297; March/April, 1954.] Formula are derived for Bessel functions of the third kind.

517.6 2995
On the Computation of Oscillatory Integrals—Y. L. Luke. (*Proc. Camb. Phil. Soc.*, vol. 50 part 2, pp. 269–277; April, 1954.) Integrals of the type $\int_a^b f(y)e^{i\lambda v}dy$ are considered. Formulas are presented which enable the integral to be easily evaluated for any λ when $f(y)$ is tabulated.

MEASUREMENTS AND TEST GEAR

621.317:621.396.828 2996
Measurements and Standards for Control of Spurious Radiation—Bailey. (*See* 3026).

621.317.3:621.396.822 2997
Errors in Power Spectra Due to Finite Sample—L. M. Spetner. (*Jour. Appl. Phys.*, vol. 25, pp. 653–659; May, 1954.) "An estimate is made of the accuracy with which the 'spectrum' of a finite sample of Gaussian noise approximates the spectrum of the original noise from which the sample was taken. The sample is considered to be spliced to form an endless loop, thereby making it a periodic function. The effect of this splice is also discussed."

621.317.326.029.63 2998
A Theory of the Diode Voltmeter for Measurement of Pulse Power at Pulsed High Frequency in the Decimetre-Wave Range—L. Kühn. (*Frequenz*, vol. 8, pp. 103–112; April, 1954.) Based on an analysis for continuous waves, the value of load resistance for equilibrium in a peak-voltmeter circuit with pulsed operation is determined. By considering an "ideal" capacitor charging characteristic which takes account of the discharge between pulses, the analysis is extended to determine charging time. The degree of rectification is almost completely dependent on the value of the parallel load resistance, which should be of the order of $10^{10}\Omega$. The problem of insulating circuit components is briefly considered.

621.317.336:621.317.755 2999
Indication and Measurement of the Argument and Modulus of an Impedance by means of the Cathode-Ray Oscilloscope—V. Popovic. (*Onde élect.*, vol. 34, pp. 376–380; April, 1954.) A simple circuit is described by means of which a direct reading in polar co-ordinates, accurate to within about 1 per cent, is obtained on the screen. The method is applicable over a wide range of frequencies. Experimental results support the theory given.

621.317.34.029.64+621.372.2 3000
An Experimental Investigation of the Single-Wire Transmission Line—Roberts. (*See* 2849.)

621.317.35:621.311.62 3001
Progress in the Measurement of Mains Harmonics—W. Grunert. (*Arch. tech. Messen*, no. 219, pp. 77–80; April, 1954.) The application of dynamometer and resonance methods to measurement of odd harmonics up to the 27th is briefly discussed. The errors in the latter method caused by a strong fundamental, by mains frequency variations and by interference are considered and formulas for estimating them are given.

621.317.4 3002
Recording Fluxmeter—R. I. Berge and C. A. Guderjahn. (*Electronics*, vol. 27, pp. 147–149; July, 1954.) An arrangement for the rapid plotting of B/H curves of ring specimens is described. The ring has two windings, one carrying a variable current supply and the other acting as search coil. Movement of the recorder pen in one direction is made proportional to the supply current and in the perpendicular direction to the flux in the specimen; a galvanometer with photocell bridge and amplifier is used. The over-all error can be kept within 0.5 per cent ±25 interlinkages/minute.

621.317.7 3003
Inductance and Dynamic Resistance Meter—G. G. Johnstone. (*Wireless World*, Aug. 1954, vol. 60, pp. 377–381; August, 1954.) An instrument developed for comparing the performances of inductors used in medium-wave receivers is described. The principle used is to feed a current of known value from an effectively infinite-impedance source to the tuned circuit comprising the inductor and tuning capacitor in parallel. A pentode tube is used as the constant-current generator.

621.317.715.044.6:538.12 3004
An Application of Conformal Transformation to the Investigation of the Magnetic Field between Galvanometer Pole-Pieces—W. Snowdon and N. Davy. (*Brit. Jour. Appl. Phys.*, vol. 5, pp. 146–151; April, 1954.) An accurate calculation is made of the magnetic potential, the associated stream function and the strength of the magnetic field along the boundary and axes of symmetry of a figure which represents the cross section of the two pole-pieces and central core of a galvanometer. The field strength between the core and pole pieces is also determined. An extension of the Schwarz-Christoffel transformation is used. Results are tabulated and shown graphically.

621.317.72/.73 3005
Electronic Test Meter—M. G. Scroggie. (*Wireless World*, vol. 60, pp. 371–376; August, 1954.) An instrument for the measurement of direct and alternating voltage and resistance is based on a combination of two tube-voltmeter designs previously described (1041 and 1671 of 1952).

621.317.723 3006
Stabilization of the Response of a Tetrode Electrometer Valve with respect to Variation of the Heating Voltage by Negative Feedback on the Cathode—D. Malé. [*Compt. Rend. Acad. Sci. (Paris)* vol. 238, pp. 1977–1979; May 17, 1954.] By connecting a resistance of appropriate magnitude in the cathode lead and tapping it off a suitable point on a voltage divider, the potential difference between the cathode and the control grid is automatically varied with variation of the heating voltage so as to maintain constant galvanometer current.

621.317.738 3007
An Instrument for Direct Measurements of Capacitance and Power Factor—J. C. Bhattacharya. (*Indian Jour. Phys.*, vol. 28, pp. 75–84; February, 1954.) Capacitances up to 0.02 μF are measured by means of an instrument whose operation is based on the variation of the frequency and Q value of a tuned circuit when the capacitance is connected across it.

621.372.8.001.4:621.372.56 3008
Precision [rectangular-] Waveguide Attenuator—W. W. H. Clarke and C. W. K. Maysmor. (*A.T.E. Jour.*, vol. 10, pp. 114–119; April, 1954.) The attenuating element is a resistive vane which is moved across the guide by a micrometer head, and which gives up to 60 db attenuation. The vane is of glass coated with a thin evaporated nichrome film, and protected by a MgF_2 layer. The ends are tapered for a distance of half the mean guide wavelength for the band concerned. Repeated calibration shows reasonable stability of readings.

621.397.6.001.4 3009
Synthetic-Pattern Generator for the Solution of Certain Instrumentation Problems in Television—R. C. Webb. (*RCA Rev.*, vol. 15, pp. 187–203; June, 1954.) Detailed description of a generator capable of providing a comprehensive range of test patterns for television testing.

621.397.62.001.4:621.397.8 3010
Factors affecting the Correlation of TV Picture Quality between Field and Laboratory Signals—B. Amos and C. Quirk. (*Trans. I.R.E.*, No. PGBTR-6, pp. 2–16; April, 1954.) Discussion of test procedures in relation to receiver design and in particular to the mass production of receivers with uniformly good performance under widely varying conditions. Over-all gain, selectivity, phase and amplitude response, and rise time of the video circuits, are considered. Conveyor-system test arrangements are described and illustrations of test patterns and cro traces are shown.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

534.15 3011
An Acoustic Method for the Measurement of Vibration Amplitudes—W. Koidan. (*Jour. Acous. Soc. Amer.*, vol. 26, pp. 428–433; May, 1954.) "By means of a modification of the pistonphone, a technique has been developed to obtain the response of a capacitance type vibration pickup probe using the reciprocity calibration of a condenser microphone as a reference."

534.321.9:621.375.2 3012
Distributed Transducer for Ultrasonic Power—J. Rabinow and M. Apstein. (*Electronics*, vol. 27, pp. 160–162; July, 1954.) A device for generating high-amplitude ultrasonic oscillations comprises a distributed amplifier with a magnetostrictive bar associated with the output. The electrical delay line is designed so that the velocity of propagation in it matches the velocity of the ultrasonic waves set up in the bar. Possible applications are indicated.

621.384.611 3013
Microtrons (Electron Cyclotrons) for X and K Band Operation—H. F. Kaiser (*Jour. Frank. Inst.*, vol. 257, pp. 89–108; February, 1954.) Design principles are discussed, and performance details given for two models operating in the 3-cm band.

621.384.612 3014
The Effect of Field Bump in a Strong-Focusing Synchrotron—K. Huke and G. Iwata. [*Jour. Phys. Soc. (Japan)*, vol. 9, pp. 297–298; March/April, 1954.] Calculation of the effect due to the field bump and the form factor.

621.384.622 3015
Transverse Motion of an Electron in a Constant—Wave-Speed Section of a Linear Accelerator—I. Jacobs and E. S. Akeley. (*Jour. Appl. Phys.*, vol. 25. pp. 572–576; May, 1954.)

PROPAGATION OF WAVES

538.566.029.61:535.22 3016
Investigation of a New Form of Microwave Interferometer for Determining the Velocity of Electromagnetic Waves—K. D. Froome. (*Proc. Roy. Soc. A*, vol. 223, pp. 195–215; April 22, 1954.) A sensitive symmetrical interferometer using two fixed horn radiators, approximately 8 m apart, and a carriage-mounted two-horn detector and indicator is described and its application to the measurement of wavelength, and hence phase velocity, is discussed. Results of preliminary measurements at 24·005 kmc give $c_0 = 299793.0 \pm 0.3$ km, the error limit given representing the standard deviation of single observations in statistical combination with systematic errors due to scattering etc. The correction of errors due to diffraction and distortion of the wavefronts was investigated in detail.

621.396.11 3017
Coupled Forms of the Differential Equations governing Radio Propagation in the Ionosphere—P. C. Clemmow and J. Heading. (*Proc. Camb. Phil. Soc.*, vol. 50, part 2, pp. 319–333; April, 1954.) The equations for oblique propagation in a horizontally stratified ionosphere with an oblique magnetic field are put in a form suitable for solution by successive approximations. In the general case this "coupled" form consists of four first-order differential equations, each associated with one characteristic wave. In the special cases of (a) horizontal magnetic field, and (b) plane of incidence perpendicular to magnetic meridian, two second-order coupled equations of a particular type can be derived, each associated with a pair of corresponding upgoing and downgoing characteristic waves. These latter equations include those given previously by Försterling (1884 of 1942) for vertical incidence. Cases in which there is no coupling are briefly considered from the point of view of the first-order equations. The general formulation provides a basis for assessing the validity of the standard "ray" approximation, alternative to that developed by Booker (3277 of 1936), and brings out the nature of its breakdown in coupling and reflection regions.

621.396.11:551.510.5 3018
N.B.S. Research in Radio Propagation—(*Tech. News Bull. Nat. Bur. Stand.*, vol. 38, pp. 49–59; April, 1954.) A brief review of the work of the Central Radio Propagation Laboratory, which has four main divisions: (a) the Ionosphere Research Laboratory, whose investigations include solar rf radiation, upper-atmospheric winds, etc; (b) the Systems Research Laboratory, which is concerned with tropospheric propagation and with research on the best use of the frequency spectrum; (c) the Measurements Standards Laboratory, which is responsible for the standard-frequency broadcasts from WWV and WWVH; (d) the propagation prediction services.

621.396.11.029.55:551.510.535 3019
Sweep Frequency Backscatter—Some Observations and Deductions—R. Silberstein. (TRANS. I.R.E., vol. AP-2, pp. 56–63: April, 1954.) Detailed analysis of records taken during two periods in December 1952 and January 1953. "In one group of observations over a 1150-km path on three undisturbed days, the values of F_2-layer maximum usable frequency scaled from midpoint vertical-incidence ionospheric records and those determined by backscatter delay assuming ground scatter agreed almost within experimental error In another three-day group characterized by a low-latitude ionospheric disturbance with low geomagnetic K indexes but considerable sporadic E activity, values of muf determined from scatter were much too high under the ground-scatter assumption, errors of about 30 per cent being not uncommon." The distant scatter in this case appears to originate at the E layer.

621.396.8 3020
The Statistical Distribution of Field Strength of Radio Signals propagated in the Troposphere and in the Ionosphere—L. Sacco.

(*Poste e Telecommunicazioni*, vol. 22, pp. 55–72; February, 1954.) An examination is made of the types of distribution in time which best represent the fluctuations of radio signals traversing the troposphere and the ionosphere. The distributions considered are (a) normal, (b) log normal, (c) fluctuations due to interference (Rayleigh distribution), and (d) fluctuations due to polarization variations. Gamma and phase distribution are discussed and the concept of correlation is introduced.

621.396.812.3.029.55:551.510.535 3021

Ionosphere and Nature of Fading Patterns of Received Radio Signals—R. N. Singh. (*Indian Jour. Phys.*, vol. 28, pp. 109–118; March, 1954.) Records of reception of sw transmissions from Indian and other stations are analyzed and compared with theoretical curves derived from Rayleigh's probability formula. A study is made of (a) variations of the nature of fading with time of day for a given signal frequency and transmitter distance, and (b) variation of the nature of fading with transmission distance for a given frequency at different times of day. The results indicate that the fading pattern is at least partially determined by the electron density in the region of the ionosphere at which the received waves are reflected.

RECEPTION

621.396.621:621.314.7 3022

An Experimental Transistor Personal Broadcast Receiver—L. E. Barton. (PROC. I.R.E., vol. 42, pp. 1062–1066; July, 1954.) See 2206 of July.

621.396.621:621.385.2 3023

Vacuum-Tube Detector and Converter for Microwaves using Large Electron Transit Angles—Bronwell, Wang, Nitz, May and Wachowski. (*See* 3096.)

621.396.621:621.396.822:621.372.54 3024

The Detection of a Sine Wave in the Presence of Noise by the Use of a Nonlinear Filter—T. G. Slattery. (PROC. I.R.E., vol. 42, p. 1124; July, 1954.) Discussion on 206 of 1953.

621.396.82:621.397.62 3025

Spurious Radiation from TV Receivers—Chittick. (*See* 3026).

621.396.828:621.317 3026

Measurements and Standards for Control of Spurious Radiation—S. L. Bailey. (TRANS. I.R.E., No. PGBTR-6, pp. 59–64; April, 1954.) Critical discussion of existing FCC regulations relating to interfering radiation. Specifications based on attenuation below unmodulated-carrier level are inconsistent with field-strength specifications applying to systems intended to be nonradiating. Two examples illustrate practical aspects of accurate field-strength measurement.

STATIONS AND COMMUNICATION SYSTEMS

621.376:621.396.65 3027

Comparison of Code and Classical Types of Modulation—J. P. Vasseur. (*Ann. Radioélect.*, vol. 9, pp. 137–149; April, 1954.) Various known systems, including pcm and delta modulation with single and double integration, are reviewed with the object of determining which is most suitable in given circumstances. Use of the code systems may enable very difficult links to be operated satisfactorily, but in most cases carrier-current, pam or ppm systems afford satisfactory operation with greater economy, both financially and from the point of view of bandwidth.

621.376.2 3028

A General Solution of the Two-Frequency Modulation Product Problem: Part 2—Tables of the Functions $A_{mn}(h,k)$—R. L. Sternberg. (*Jour. Math. Phys.*, vol. 33, pp. 68–79; April, 1954.) Methods and tables are presented for evaluating functions discussed in Part 1 [2212 of July (Sternberg and Kaufman)].

621.376.5:621.396.41 3029

"Elastic" Pulse Modulation—F. J. Tischer. (*Fernmeldetech. Z.*, vol. 7, pp. 187–189; April, 1954.) The term "elastic pulse modulation" is used to designate a particular system in which quantization of the signal is performed at both transmitter and receiver. The operation is described by comparing it with an ordinary ppm system in which the transmitter pulses are produced by superposing the modulation voltage on a sawtooth voltage, while at the receiver the incoming pulses are used to gate a tube which has a synchronous sawtooth voltage applied to a second grid. In the new system the sawtooth is replaced by a stepped waveform, with the object of reducing the effects of noise in the transmission channel. From the statistical distribution of noise amplitude it is possible to determine, for given channel width, pulse shape and number of steps, the fraction of noise which will appear in the output. The method greatly reduces crosstalk in multichannel systems.

621.376.56 3030

Conjugate Pulse Modulation with Linear Detection—J. A. Ville. (*Onde élect.*, vol. 34, pp. 372–375; April, 1954.) An analysis is made of pulse-code systems using equispaced pulses with on-off modulation. The pulse groups are decoded by simple filtering. A determination is made of the best law of coding to be used at the transmitter in conjunction with a receiving filter of known characteristics.

621.39.001.11 3031

Symposium on Statistical Methods in Communication Engineering—(*Trans. I.R.E.*, No. PGIT-3, pp. 1–159; March, 1954.) The text is given of the following papers presented at a symposium at Berkeley, Calif., in August, 1953:

A. J. F. Siegert, "Passage of Stationary Processes through Linear and Nonlinear Devices," pp. 4–25.
D. Middleton, "Statistical Theory of Signal Detection," pp. 26–51.
R. C. Davis, "Detectability of Random Signals in the Presence of Noise," pp. 52–62.
B. Gold and G. O. Young, "The Response of Linear Systems to Non-Gaussian Noise," pp. 63–67.
D. Slepian, "Estimation of Signal Parameters in the Presence of Noise," pp. 68–89.
D. C. Youla, "The Use of the Method of Maximum Likelihood in Extimating Continuous-Modulated Intelligence which has been Corrupted by Noise," pp. 90–105.
R. Deutsch, "Detection of Modulated Noise-Like Signals," pp. 106–122.
O. J. M. Smith, "Statistically Almost Optimum Nonlinear Network Design," (abstract only), p. 123.
B. Mandelbrot, "Simple Games of Strategy occurring in Communication through Natural Languages," pp. 124–137.
A. E. Laemmel, "Application of Linear Graphs to Communication Problems," (abstract only), p. 138.
N. M. Blachman, "Minimum-Cost Encoding of Information," pp. 139–149.
W. P. Caywood, Jr., R. C. Lyman, and W. M Kaufman, "Generalized Servomechanism Evaluation," (abstract only), p. 150.
A. J. Mallinckrodt and T. E. Sollenberger, "Optimum Pulse-Time Determination," pp. 151–159.

621.39.001.11 3032

Information Theory and Knowledge—D. K. C. MacDonald. (*Jour. Appl. Phys.*, vol. 25, pp. 619–622; May, 1954.) "Distinction between data transmission from one person (or point) to another and the inception of new knowledge is analyzed in relation to information theory."

621.39.001.11 3033

Negentropy and Information in Telecommunications, Writing, and Reading—L. Brillouin. (*Jour. Appl. Phys.*, vol. 25, pp. 595–599; May, 1954.) The Shannon formula for the capacity of a channel with noise is derived from the negentropy principle of information (794 of March) by thermodynamic considerations. The problem of writing and reading is examined; the source of light needed for reading provides the negentropy which is changed into information by the reader. The negentropy connected with organization is also discussed.

621.395.44 3034

The Philips Short-Haul Carrier-Telephone System STR 112—(*Commun. News*, vol. 14, pp. 77–128; March/April, 1954.) A series of articles describing the system, which was designed primarily to the requirements of the Dutch PTT, and which is intermediate between the voice-frequency system and the standard 4-kc-spaced system in present use. The STR 112 is a ssb system in which the carriers are spaced at 6-kc intervals, and the 2-kc spaces between the 4-kc speech channels are used for signaling. Equipment for a complete channel is contained in a single sealed unit.

621.396 3035

High-Frequency Communications—(*Elec. Eng.*, vol. 26, pp. 223–278; June, 1954.) Practically the whole of this issue is devoted to a review of existing practice in the field of radio communication at frequencies mainly in the range 3–30 mc and to consideration of problems in this field still awaiting solution. The individual papers include:

D. Scott, "Commonwealth Radio Communication Services," pp. 224–231.
A. W. Cole, "H.F. Communication Systems," pp. 232–236.
W. J. Morcom, "A High-Power Communication Transmitter," pp. 237–241.
F. W. J. Sainsbury, "New Developments in H.F. Receivers," pp. 241–249.
D. J. Spooner, "1.5 to 30 Mc/s at Sea," pp. 250–253.
G. L. Warner, "H.F. Airborne Communication Equipment," pp. 253–255.
D. H. C. Scholes, "Radio Teleprinter Systems," pp. 256–259.
A. G. Williamson, "Developments in Frequency-Shift Keying and Radio Teleprinter Systems," pp. 260–265.
R. Terlecki, "A Narrow-Band Frequency-Shift Telegraph System," pp. 265–267.
S. C. Howard, "Frequency-Shift Diplex," pp. 268–271.
J. R. Humphreys, "Naval Low-Power M.F.–H.F. Communications," pp. 272–274.
C. Gillam, "High-Power Aerial Switching," pp. 274–278.

621.396.1 3036

Message Error in Diversity Frequency-Shift Reception—G. F. Montgomery. (PROC. I.R.E., vol. 42, pp. 1184–1187; July, 1954.) For a keyed signal, errors due to carrier fading can be specified in terms of the binary error. The binary error is calculated for several diversity selection methods, assuming narrow-band transmission, fluctuation noise interference, and a Rayleigh distribution of carrier amplitudes. The results are used to determine what reduction of transmitted carrier power can be permitted on substituting a given diversity receiving system for a single receiver.

621.396.3:621.376.3 3037

Design Considerations for Frequency-Shift-Keyed Circuits—W. Lyons. (*RCA Rev.*, vol. 15, pp. 230–237; June, 1954.) Analysis indicates that a considerable improvement of signal/noise ratio is obtained by using sine-wave instead of rectangular-pulse keying. Unity ratio of frequency deviation to modulation frequency gives best results. A particular keying speed giving optimum use of the frequency band is determined.

621.396.4 3038

Four-Channel FSK adds Radio Circuits—C. Buff. (*Electronics*, vol. 27, pp. 156–159; July, 1954.) A new frequency-shift keyer capable of adapting any class-C radiotelegraph

transmitter to multichannel operation is described. Four channels, including keying sidebands, are accommodated in an over-all bandwidth of 3.85 kc. Heterodyning is used, in place of the more usual frequency multiplication, to obtain carrier frequencies in the range 4–24 mc.

621.396.49:621.397.5 3039
Considerations regarding the Rational Development of Television from the Economic Point of View—Schröter. (*See* 3051.)

621.396.5:621.396.65 3040
Subscribers' Radio Telephone—M. O. Felix and C. B. Wooster. (*A.T.E. Jour.*, vol. 10, pp. 93–103; April, 1954.) Radio links between telephone subscribers and the exchange are required where line construction is impracticable or too costly. The design is considered of signaling and terminating units such that a radio link can be interposed in a standard line circuit without modification to the switching equipment or operating procedure. A comprehensive series of units is described, including battery-operated subscribers' equipment. Carrier suppression with frequency shift is the signaling method chosen.

621.396.65.029.63:621.372.55 3041
Use of Pulse Systems for the Simultaneous Transmission of Telephony, Telegraphy and Radio Programmes—G. X. Potier. (*Onde élect.*, vol. 34, pp. 347–355; April, 1954.) Modifications necessary for adapting a ppm multiplex telephony system to handle program and telegraphy signals also are discussed. The pulse repetition rate must be increased, typically by a factor of 2, to accommodate the program signals; the multiplication may be performed either before or after modulation, but distortion considerations indicate that only the latter method is practical. Equipment operating with a carrier frequency of 1.7–2.3 kmc is described in which two telephony channels are conveniently replaceable by one program channel.

621.396.721.029.51/.53:621.396.677 3042
The Problem of the Long-Wave Radio Relay—R. Kümmich. (*Frequenz*, vol. 8, pp. 112–122 and 159–166; April/May, 1954.) A detailed account is given of practical difficulties involved in establishing relay stations designed during wartime to give false df bearings. The main problem was to isolate the receiver from the radiation of the associated transmitter. In designing a directive receiving antenna for a transmitter-receiver separation of 2 km a reduction factor of $1:10^4$ was aimed at but never achieved despite protracted investigations at two sites in North Jutland and in Holstein. This was due primarily to the variation of site error with frequency. Night effect and the effect of ground inhomogeneities, pickup via the hf connecting cable and antenna movement on cardioid and figure-of-eight polar diagrams are discussed and refinements of loop and Adcock systems are described. A description is given of equipment operated, with details of its performance at 371 kc with bandwidth about 10 kc. Using an Adcock receiving system 10 km from a 1.5-kw transmitter a reduction factor of 1:2000 was obtained.

621.398:621.396.712:621.376.3 3043
A Radio-Relay Remote-Control System for FM Broadcast Stations—T. R. Humphrey. (*Elec. Eng.*, vol. 73, pp. 327–330; April, 1954.) Station WVBT of the Rural Radio Network, operating on 91.5 mc, is controlled from Station WVCN, 80 miles away, by means of modulation signals in the 19–24-kc band. Information on the power-amplifier anode voltage and current, rf output voltage and ac line-voltage is returned via a 30-kc modulation of the WVBT transmission. Block diagrams of the system are given. To be published also in *Trans. AIEE, Communication and Electronics;* 1954.

SUBSIDIARY APPARATUS

621–526 3044
Analysis of Nonlinear Servos by Phase-Plane-Delta Method—R. N. Buland. (*Jour. Frank. Inst.*, vol. 257, pp. 37–48; January, 1954.)

621.311.6 3045
Series Capacitors multiply Battery Voltage—J. M. Reed. (*Electronics*, vol. 27, pp. 182, 188; July, 1954.) The voltage obtainable by charging capacitors in parallel and discharging them in series is increased by using a set of capacitors of graded capacitance values and charging the first by the battery only, the second by the battery in series with the first capacitor, and so on. The method has been used to provide 6.4 kv, using six capacitors with a 100v battery.

621.316.721/.722].078.3 3046
Voltage and Current Stabilizers for Laboratory Use—J. Mevel. (*Jour. Phys. Radium*, vol. 15, pp. 290–298; April, 1954.) A systematic exposition of the principal types of stabilizer, divided into three main classes, (a) direct, (b) feedback, and (c) other types. 151 references.

621.316.722.025 3047
A.C. Voltage Stabilizers—W. M. Dalton. (*Elec. Eng.*, vol. 26, pp. 310–313; July, 1954.) "The operation of the output stage of the tube-operated ac mains stabilizer is analyzed with the aid of load lines and the relative advantages of various modes of operation are briefly discussed."

621.352.7:537.311.33 3048
Semiconductors as Solid Electrolytes in Electrochemical Systems—K. Lehovec and J. Broder. (*Jour. Electrochem. Soc.*, vol. 101, pp. 208–209; April, 1954.) A "single-shot" battery and a long-life cell are described and the physico-chemical principles of operation outlined. The former, an Ag-AgI-Ag_2S-S cell, gave an open-circuit voltage of ~0.2v at 200 degrees C. The short-circuit current was ~0.18 A/cm². The latter, an Ag-AgI-I(+C) cell with an emf of ~0.7v at room temperature, can deliver a current of several hundred μA/cm².

TELEVISION AND PHOTOTELEGRAPHY

621.397.2:621.397.743 3049
Networks for Theater Television—F. A. Cowan. (*Jour. Soc. Mot. Pict. Telev. Eng.*, vol. 62, part I, pp. 306–313; April, 1954.) An illustrated account of the 4-kmc-band radio relay system and the L1 and L3 vestigial-sideband coaxial-cable systems, with carrier frequencies of 311 kc and 4.139 mc respectively. The systems serve 204 stations in 130 U.S. cities.

621.397.5:621.395.625.3 3050
Recording Television and Sound Signals on Magnetic Tape—(*Elec. Eng.*, vol. 26, pp. 292–295; July, 1954.) See 2517 of August (Olson *et al.*).

621.397.5:621.396.49 3051
Considerations regarding the Rational Development of Television from the Economic Point of View—F. Schröter. (*Bull. schweiz. elektrotech. Ver.*, vol. 45, pp. 237–243; April 17, 1954.) A two-stage plan is proposed to facilitate the development of television in countries with small populations or low purchasing power. A network of usw transmitters would be established, having bandwidths large enough for television, but operated in the first stage as sound transmitters only. A number of programs would be available and would be sampled at picture-line frequency, a composite signal similar to the television video signal being radiated. At the receiver the desired program would be selected by using the synchronizing signal, after passage through a delay line, as a gating pulse for a selector tube. The zero-modulation output level of the transmitter for sound would coincide with the black level for the television transmission in the second stage. The network could be used for sound transmissions during television pauses in the second stage.

621.397.5:621.396.712.3 3052
C.B.S. Television City Technical Facilities—H. A. Chinn, R. S. O'Brien, R. B. Monroe, and P. E. Fish. (Proc. I.R.E., vol. 42, pp. 1067–1091; July, 1954.) A detailed account is given of the CBS television headquarters at Hollywood, planned eventually to occupy the whole of a 25-acre site. At the interim stage reached in 1953, four studios, two with accommodation for an audience and each with an area $>\frac{1}{4}$ acre, were completed; the design allows for 24 studio units in all, and is specially planned to facilitate expansion. A number of new techniques have been introduced.

621.397.5(083.7) 3053
I.R.E. Standards on Television: Methods of Measurement of Aspect Ratio and Geometric Distortion, 1954—(Proc. I.R.E., vol. 42, pp. 1098–1103; July, 1954.) Standard 54 I.R.E. 23S1. Correction, *ibid.*, vol. 42, p. 1314; August, 1954.

621.397.5(083.74) (931) 3054
Choosing a Television Standard—(*Wireless World*, vol. 60, pp. 401–402; August, 1954.) See 2235 of July.

621.397.6(001.4) 3055
Synthetic-Pattern Generator for the Solution of Certain Instrumentation Problems in Television—Webb. (*See* 3009.)

621.397.61:621.375.2.029.63:621.385.029.63 3056
A U.H.F. Transmitter employing Klystron Power Amplifiers—W. H. Sayer. (*Elec. Eng.*, vol. 73, pp. 354–359; April, 1954.) Applications are discussed of the 3K20,OOOLK klystron in both narrow- and wide-band power amplifiers operating in the 720–890-mc frequency band. The circuit efficiency of a klystron output circuit in a 5-kw television transmitter is 97.9 per cent compared with 73 per cent for the 4×1500 tetrode used as modulated amplifier delivering 120w peak. The demountable cavities allow tuning without changing the gap spacing of the drift tubes. Two other similar klystrons will cover the 470–720-mc frequency band. To be published also in *Trans. AIEE, Communication and Electronics;* 1954.

621.397.611.2 3057
Electronic Light-Trace Marker, a New Apparatus for Television Studios and Outside-Broadcast Equipment—R. Sondermeyer. (*Funk u. Ton*, vol. 8, pp. 179–186; April, 1954.) A portable unit has been developed for superimposing a light trace of triangular or other form on a picture during transmission. The trace can be adjusted to any position in the frame by means of a hand-operated control. A detailed description is given of the derivation of the pulse train forming the marker trace from horizontal and vertical blanking pulses.

621.397.611.2:778.5 3058
Continuous All-Electronic Scanner for 16 mm Color Motion-Picture Film—V. Graziano and K. Schlesinger. [*Jour. Soc. Mot. Pict. Telev. Eng.*, vol. 62, part I, pp. 294–305; April, 1954. *Proc. NEC* (*Chicago*), vol. 9, pp. 370–381; 1953.] The Motorola flying-spot system and equipment are described. The kinescope scans the odd-numbered film frames three times and the even-numbered frames twice, thus converting the 24 frames/seconds to 60 television fields/seconds. A signal derived from the leading edges of the sprocket perforations is used to control the position of the scan on the tube screen and is also used for continuous correction of film shrinkage. The anastigmatic deflection yoke and the adjustable gamma control are also briefly described.

621.397.62 3059
Printed Circuit Television Set—G. Bovill. (*Wireless World*, vol. 60, pp. 363–366; August, 1954.) Use of printed-circuit technique may be more useful for television than for ordinary sound receivers, because the main design features are more likely to remain stable. An illustrated description is given of an 819-line

receiver of French manufacture, for single-channel operation on the 174–185-mc band, in which the wiring, coils, and smaller capacitors are printed.

621.397.62 3060

Combined 625-819 Receiver—R. Gondry. (*Télévision*, Nos. 42 and 43, pp. 85–91 and 122–127; March/May, 1954.) A multichannel receiver designed particularly for the transmissions from Baden-Baden, Strasbourg, and Paris in the upper vhf band. A detailed description of the receiver construction is given with diagrams of circuit arrangement, chassis dimensions, and coil data. Alignment procedure and switching arrangements are described. Operating data are given and characteristic oscillograms are shown.

621.397.62:621.386.842 3061

A Solid-State Image Intensifier—R. K. Orthuber and L. R. Ullery. (*Jour. Opt. Soc. Amer.*, vol. 44, pp. 297–299; April, 1954.) A method of image intensification which avoids the need for an evacuated envelope uses a layer of a photoconductor such as *CdS* in contact with a layer of electroluminescent phosphor. In an example, the brightness gain is of the order of two; this is in satisfactory agreement with the theoretical value. The construction of the image intensifier is described and its application to the intensification of projected television pictures is mentioned.

621.397.62:621.396.82 3062

Spurious Radiation from TV Receivers—K. A. Chittick. (*Trans. I.R.E.*, No. PGBTR-6, pp. 71–77; April, 1954.) Recent investigations of radio interference and television-receiver oscillation are reported. Standards proposed for broadcast-band interference limit the field generated by the television-receiver sweep-oscillator to 100 μV/m at 15 feet, and the voltage on a power line from the same source to 100 μV between line and earth. The field strength required at the television antenna to give a reasonable picture signal is discussed. Possible interference to different services due to local-oscillator radiation throughout the television bands is considered; the present standard tolerance for vhf, particularly channels 4 and 7, should be reduced by half to provide a greater safety factor for the omnidirectional-range services and amateur stations; the limit of 500 μV/m for uhf seems satisfactory.

621.397.62.001.4:621.397.8 3063

Factors affecting the Correlation of TV Picture Quality between Field and Laboratory Signals—Amos and Quirk. (*See* 3010.)

621.397.621:621.318.4 3064

High-Efficiency, Low-Copper Sweep Yokes with Balanced Transient Response—C. E. Torsch. (*Trans. I.R.E.*, No. PGBTR-6, pp. 17–32: April, 1954.) An improved design of deflection-coil assembly is described. It gives a sweep angle of 70 degrees and is no larger than an earlier 50 degree-sweep unit. It uses the shorter coils for the horizontal sweep, one of this pair of coils being wound in reverse; if the pair is connected in parallel no balancing capacitor is required. Coils have interlocking flared ends, a design permitting reduction of the length of the assembly. Advantages of the coil reversal are discussed and improved techniques of winding and insulation are noted. See also 2916 of 1950.

621.397.621:621.318.4:535.623 3065

Extension of the Balanced-Transient Response Principle to Color TV Yokes—C. E. Torsch. (*Trans. I.R.E.*, No. PGBTR-6, pp. 33–35; April, 1954.) The principles of reverse winding and parallel connection of the coil halves (3064 above) are applied to vertical-sweep windings. Connections are arranged so that the high-potential turns of one pair of coils are adjacent to the low-potential turns of the quadrature coils.

621.397.7:621.375.2.029.62 3066

Line Amplifiers for Community Television Systems—K. A. Simons, D. Kirk, and H. J. Arbeiter. (*Elect. Eng. N.Y.*, vol. 73, pp. 337–342; April, 1954.) Discussion of the relative technical and economic merits of (a) single-channel, (b) cascade, and (c) distributed broadband amplifiers in a three-channel and a seven-channel vhf television system. Results show that type (a) is to be preferred in large systems, type (b) in small systems with short cable runs and few repeaters. To be published also in *Trans. AIEE, Communication and Electronics;* 1954.

621.397.813 3067

Method of Equalizing Television Transmissions by Deliberate Introduction of Echoes—J. Müller. (*Fernmeldetech. Z.*, vol. 7, pp. 199–204; April, 1954.) A circuit is described in which delay lines are used to introduce short-delay echoes so as to produce time equalization of the transient response up to 0.5 μs. The improvements effected are illustrated by oscillograms.

621.397.82 3068

Television Receiver Interference Industry Record to Date—L. M. Clement. (*Trans. I.R.E.*, No. PGBTR-6, pp. 50–58; April, 1954.) Investigations made by the U.S. Radio Manufacturers Association in conjunction with other bodies are reported. Interference from police transmitters in television receivers using the standard 41.25-mc if is noted. A survey has been made of the progress of this standardization. Information based on answers to questionnaires is presented, relating to the use made of field-test facilities and to measurements of oscillator radiation from television and fm receivers.

TUBES AND THERMIONICS

621.314.63:546.289 3069

Capacitance Measurements on Alloyed Germanium-Indium Rectifiers—H. L. Rath. (*Naturwissenschaften*, vol. 41, pp. 161–162; April, 1954.) Independent determinations of the impurity concentration based on (a) conductivity, and (b) capacitance give results in agreement with one another.

621.314.632:546.289 3070

Observations on a Form of Breakdown in Germanium Diodes—A. Lempicki and C. Wood. (*Proc. Phys. Soc.*, vol. 67, pp. 328–337; April 1, 1954.) The I/V characteristic of the Ge diode is shown to depend on the waveform applied. The critical barrier temperature at which the breakdown observed by Billig (437 of 1953) occurs and the barrier radius are obtained by postulating a simple model for the contact. The possibility that the breakdown is due to thermal instability is discussed.

621.314.7 3071

New Advances in the Junction Transistor—E. W. Herold. (*Brit. Jour. Appl. Phys.*, vol. 5, pp. 115–126; April, 1954.) A survey of recent work in the U.S.A. on alloy junction transistors, their geometry, characteristics and applications. See also Mueller and Pankove (1955 of June).

621.314.7:546.28 3072

Silicon Transistor—W. A. Adcock, M. E. Jones, J. W. Thornhill, and E. D. Jackson. (Proc. I.R.E., vol. 42, p. 1192; July, 1954.) Measurements are reported on *n-p-n* transistors made from single crystals of Si, with base layers about 0.0005 inches thick. The results confirm the view that Si transistors can handle more power than similar Ge units.

621.314.7:621.357 3073

Electrolytic Analog Transistor—H. Letaw, Jr. and J. Bardeen. (*Jour. Appl. Phys.*, vol. 25, pp. 600–606; May, 1954.) An operating model of a junction transistor is described which substitutes the reduced and oxidized forms of ions in solution for electrons and holes in a semiconductor. The emitter and collector are circular platinum electrodes of diameter 0.5 inch, with adjustable spacing. A mercury pool in contact with mercurous chloride and chloride ion makes a satisfactory base electrode. The cell is of lucite. Because the mobility of the ions is much lower than that of electrons and holes, flow phenomena can be studied more readily. This electrolytic transistor may also prove useful as a constant-current device.

621.314.7:621.372.5 3074

Equivalent Circuits of Transistors, and their Physical Basis—J. Malsch. (*Arch. elekt. Übertragung.*, vol. 8, pp. 179–189; April, 1954.) The form of equivalent circuit originally introduced to represent the point-contact transistor was found to require modification in the light of increased knowledge of the physics of transistors and the development of junction types. An equivalent circuit is presented which is based on diffusion-theory analysis and gives a satisfactory representation of the transistor with frequency-independent elements up to relatively high frequencies.

621.38.002.2:535.336.2 3075

Mass Spectrometry in the Electronics Industry—T. G. Polanyi. (*Sylvania Tech.*, vol. 7, pp. 51–58, 61; April, 1954.) Applications in research, development, and production, are described.

621.383:546.28 3076

A New Silicon *p-n* Junction Photocell for Converting Solar Radiation into Electrical Power—D. M. Chapin, C. S. Fuller, and G. L. Pearson. (*Jour. Appl. Phys.*, vol. 25, pp. 676–677; May, 1954.) Photons of energy 1.02 eV or more ($\lambda \leqq 1.2\,\mu$) are able to produce electron-hole pairs in Si; in the presence of a *p-n* barrier these pairs are separated and do work in an external circuit. Calculation indicates that the upper limit of efficiency of such a cell is about 22 per cent; in practical devices the efficiency is reduced by reflection of radiation at the Si surface, by recombination of electrons and holes before reaching the barrier, and by losses due to surface-layer and contact resistance. See also 2779 of September.

621.383.27 3077

Fatigue in an Electron-Multiplier Photocell—F. Lenouvel and J. Daguillon. (*Jour. Phys. Radium*, vol. 15, pp. 287–289; April, 1954.) The fatigue observed in a photomultiplier passing an anode current of 4×10^{-7}A is due to the resistance of the semiconductor film covering the electrode in the final multiplier stage. This resistance is about 200 MΩ at room temperature. The electrodes of the preceding stages do not contribute appreciably to the fatigue until the anode current reaches higher values.

621.385:538.691 3078

Electron Trajectories in Strip Beams Constrained by a Magnetic Field—J. D. Lawson. (Proc. I.R.E., vol. 42, pp. 1147–1151; July, 1954.) Analysis is presented for idealized cases. Different types of solution are obtained according as the electron gun is outside or inside the magnetic field; in the former case the nature of the flow depends also on the current density relative to a critical value. The beamwidth varies periodically along the beam in all cases.

621.385:621.318.57 3079

Development of Trochotrons—J. Björkman and L. Lindberg. [*Kungl. tek. Högsk. Handl.* (*Stockholm*), No. 80, 131 pp.; 1954.] A comprehensive account of the design, construction and applications of magnetron-type switching tubes previously described [1205 of 1948 (Alfvén & Romanus), etc.].

621.385:621.396.822 3080

The Noise of Valves—C. S. Bull. (*Brit. Jour. Appl. Phys.*, vol. 5, pp. 127–132; April, 1954.) Theory developed in 909 of March and back references is restated and extended to include the effects of screen grids and secondary-electron multiplication.

621.385.004.15 3081

Electron-Tube Life and Reliability: Part 1—

M. A. Acheson. (*Sylvania Tech.*, vol. 7, pp. 45–50; April, 1954.) "The reliability of an electron tube may be described in terms of its failure rate for a given type of failure under specified conditions of operation including environmental, mechanical, and electrical factors. When failure rates are determined by proper analysis of obtainable data, the degree of reliability that may be expected can be predicted by use of statistical procedures. In simple cases the binomial formula may be used. When larger quantities of tubes are involved, the simpler Poisson formula may be used. Applications are described in elementary terms, and several examples are given."

621.385.029.6 3082
Stored Energy and Power Flow in Electron Beams—L. R. Walker. (*Jour. Appl. Phys.*, vol. 25, pp. 615–618; May, 1954.) Analysis of the interaction between a weak traveling wave and a homogeneous beam indicates that the density of ac energy and the ac power flow in the beam depend on the manner in which the traveling wave is established.

621.385.029.6 3083
Experimental Investigation of the Operation of the Reflection Space of a Reflex Klystron—M. de Bennetot. (*Ann. Radioélect.*, vol. 9, pp. 150–162; April, 1954.) Continuation of work previously reported (3159 of 1953). The length of the drift space is determined, and the influence of the angle of incidence of the electrons in the drift space and of the reflector curvature is assessed, with and without space charge. The influence of higher-order terms in the expressions for the transit angle and current are evaluated. Measurements on a Type-KR370 tube are reported. It is concluded that the bunching efficiency cannot be found simply by calculation, but can be determined approximately by measurements.

621.385.029.6 3084
Some Recently Obtained Results on Valves for Microwave Frequencies—R. R. Warnecke. (*Ann. Radioélect.*, vol. 9, pp. 107–136; April, 1954.) A review with special attention to the magnetron-type traveling-wave amplifier and the carcinotron oscillator. 74 references. For a similar review in English see 901 of March (Warnecke and Guénard).

621.385.029.6 3085
Fundamental Phenomena in Travelling-Wave Valves—R. Warnecke, P. Guénard, and O. Doehler. (*Onde élect.*, vol. 34, No. 325, pp. 323–338; April, 1954.) A résumé is given of theory covering various types of traveling-wave tubes, including both amplifiers and oscillators. See also 901 of March (Warnecke and Guénard).

621.385.029.6 3086
Traveling-Wave Tubes—R. G. E. Hutter, S. Cutler, and H. Greenberg. (*Radio & Telev. News, Radio-Electronic Eng. Section*, vol. 51, pp. 23–25, 71; April, 1954.) The main features and principles of operation of various types are outlined.

621.385.029.6:621.372.2 3087
Investigation of Lines with Periodic Bar Structure for U.H.F. Valves—A. Leblond and G. Mourier. (*Ann. Radioélect.*, vol. 9, pp. 180–192; April, 1954.) General theory, based on that for multiconductor lines, is given for delay lines constituted by arrangements of bars; such lines have properties intermediate between those of the helix and those of the multicavity circuit. The characteristic impedance introduced by Fletcher (3275 of 1952) is used in conjunction with the transverse field distribution to calculate the energy stored in the vicinity of the bars. The circuit of the interdigital magnetron is analyzed in detail.

621.385.029.6:621.373.423 3088
The Frequency Spectrum of the Strapped Magnetron—L. Palócz. (*Acta tech. Acad. Sci. hungaricae*, vol. 9, pp. 135–147; 1954. In English.) The frequency spectrum of the unstrapped magnetron and of a double-ring strapped type are deduced from considerations of the respective equivalent circuits. A simple formula is derived for the frequency separation of the $\frac{N}{2}$, or π, mode from the adjacent $(\frac{N}{2}-1)$ mode in the strapped magnetron.

621.385.029.63/.64 3089
Bifilar Helix for Backward-Wave Oscillators—Ping King Tien. (Proc. I.R.E., vol. 42, pp. 1137–1143; July, 1954.) By applying a potential difference between the wires of the bifilar helix, a periodic beam-focusing field is provided. Measurements on an experimental tube are reported; on varying the potential difference from 0 to 2.3 kv the collector current passed through a maximum at 480v, approximately the value of the beam voltage. Oscillations at frequencies between 3 and 3.5 kmc were obtained.

621.385.029.63 3090
A Developmental Medium-Power Traveling-Wave Tube for Relay Service in the 2000-Megacycle Region—W. W. Siekanowicz [Proc. I.R.E., vol. 42, pp. 1091–1097; July, 1954. *Proc. NEC (Chicago)*, vol. 9, pp. 601–608; 1953.] The tube described has a power output of 2.5w or more, a gain of 20–30 db and a small-signal bandwidth of 500 mc. The operating voltage required is 700v. Coaxial input and output connections are provided to the helix, which has a length of about 3 inches. The design includes a conventional *U*-type permanent magnet, weighing 10–15 lb, for beam focusing.

621.385.029.63:621.397.61:621.375.2.029.63 3091
A U.H.F. Transmitter Employing Klystron Power Amplifiers—Sayer. (*See* 3056.)

621.385.029.64 3092
Nonreciprocal Loss in Traveling-Wave Tubes using Ferrite Attenuators—J. S. Cook, R. Kompfner, and H. Suhl. (Proc. I.R.E., vol. 42, pp. 1188–1189; July, 1954.) Unwanted oscillations in traveling-wave tubes with helical lines can be prevented by surrounding the helix with a ferrite cylinder with a circumferential component of magnetization, which introduces small loss in the direction of electron flow and large uniformly distributed loss in the backward direction.

621.385.029.64.001.57 3093
A Mechanical Traveling-Wave Oscillator—C. C. Cutler. (*Bell Lab. Rec.*, vol. 32, pp. 134–138; April, 1954.) A simple magneto-mechanical model is described for demonstrating the coupling between waves and electron streams such as occurs in traveling-wave tubes. The analogies discussed reveal wave-interaction characteristics which may have applications in lf transmission-line circuits.

621.385.032.216 3094
Transient Phenomenon in the Oxide Cathode Layer—T. Shindo. [*Jour. Phys. Soc. (Japan)*, vol. 9, pp. 38–43; January/February.] The theory previously given (591 of February) is extended to clarify the transient phenomenon in the building-up of the space charge near the surface of the oxide layer.

621.385.2:537.525.92 3095
Approximate Solutions of the Space-Charge Problem for Some Unusual Electrode Geometries—H. F. Ivey. (*Jour. Appl. Phys.*, vol. 25, p. 543; April, 1954.) Correction to paper abstracted in 1973 of June.

621.385.2:621.396.621 3096
Vacuum-Tube Detector and Converter for Microwaves using Large Electron Transit Angles—A. B. Bronwell, T. C. Wang, I. C. Nits, J. May, and H. Wachowski. (Proc. I.R.E., vol. 42, pp. 1117–1123; July, 1954.) An analysis is made of the motion of electrons in space-charge-limited diodes, for very large transit angles; the effect of the Maxwell-Boltzmann distribution of emission velocities is taken into account. The net effect of the microwave field is to increase the rate at which electrons are removed from the space-charge region; this explains the observed increase of anode current and the detection and conversion effects.

621.385.3:621.375.232.4.029.62:621.396.822 3097
Study of Noise Reduction by Feedback in Ultra-High Frequency Amplifiers—Glenn. (*See* 2899.)

621.385.4-71 3098
Heat Transfer from Electron Tubes at High Altitudes and High Ambient Temperatures—B. O. Buckland. (*Trans. AIEE*, vol. 72, part I, *Communication and Electronics*, pp. 698–703; 1953. Discussion, p. 703.) Results are reported of tests on a Type-6AQ5 beam power tetrode at simulated high altitudes.

621.385.832 3099
Focusing of an Electron Beam by Periodic Fields—A. M. Clogston and H. Heffner. (*Jour. Appl. Phys*, vol. 25, pp. 436–447; April, 1954.) Analysis is presented for the four cases of periodic fields of the axially symmetric and four-pole types with focusing either by electric or by magnetic fields. Under certain stated conditions parallel flow will be obtained in the beam. Actual trajectories as plotted by the analog computer are shown and charts and equations assembled to aid in the design of practical focusing systems.

621.385.832 3100
A Monoscope Tube for Computer and other Applications—J. Hartmann. (*Elec. Eng.*, vol. 73, pp. 208–212; March, 1954.) Description of the Type-K1043 monoscope tube and associated circuits. The tube is designed to resolve 30 lines of a symbol, using a second-anode supply of 1 kV and single-ended deflection.

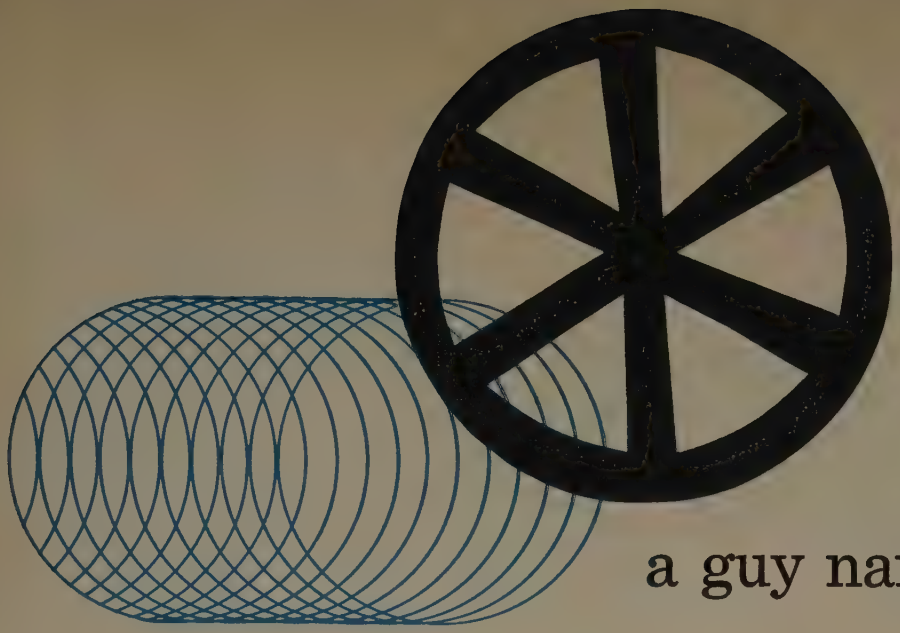
621.387:621.316.722 3101
Impedance-Frequency Variations of Glow-Discharge Voltage-Regulator Tubes—F. A. Benson and G. Mayo. (*Elec. Eng.*, vol. 26, pp. 206–210; May, 1954.) Measurements were made over the range 20 cps to 10 kc of the impedance of tubes of Types 85A1, 85A2, 7475 and S130. The tube impedance includes both resistive and reactive components, the reactance being inductive. The impedance and resistance of all tubes increase approximately by factors of 4 and 2 respectively over the range. The inductance shows a pronounced maximum at about 100 cps.

MISCELLANEOUS

621.3(083.7) 3102
I.R.E. Standards on Graphical Symbols for Electrical Diagrams, 1954—(Proc. I.R.E., vol. 42, pp. 965–1020; June, 1954.) Standard 54 I.R.E. 21S1.

621.396:061.3 3103
Convention Record of the I.R.E.—This publication, comprising ten separate parts, contains the full text or abstracts of papers presented at the I.R.E. National Convention, New York, N. Y., March 22–25, 1954.

621.396 3104
Advances in Electronics, Vol. IV [Book Review]—L. Marton (Ed.). Publishers: Academic Press, New York, N. Y., 344 pp., 1952, $8.50. (*Elec. Eng.*, vol. 26, pp. 220–221; May, 1954.) Vol. III: 2674 of 1952.



a guy named Og

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Later, your name was Watt. Steam made your kettle-lid dance...and the Industrial Revolution was on.

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Industrial Engineering Notes

FCC Actions

The Federal Communications Commission finalized, in part, its proposed rule making and amended Part 16 of its Rules Governing Land Transportation Radio Services to establish a Motor Carrier Radio Service, effective Oct. 15, 1954. The new service will replace the existing Intercity Bus and Urban Transit Radio Services. Four frequencies in the 44.46 to 44.58 mc band are allocated for exclusive use by urban transits, and ten frequencies in the 44.06 to 44.42 mc band are allocated for exclusive use by intercity buses, with 44.06, 44.10 and 44.14 mc being subject to the outcome of further proposed rule making. That part of the proposal which would have included in the Motor Carrier Radio Service the carriers of property now operating under the Highway Truck Radio Service has not been adopted. Instead, the Commission has adopted a Notice of Further Proposed Rule Making looking toward the inclusion of property carriers in the new service with eligibility limited to common carriers and possible restrictions placed on the use of radio within cities. . . . **The Federal Communications Commission amended Parts 7 and 8 so as to make available for immediate use an additional pair of mobile radiotelephone frequencies for the San Francisco-Eureka, Calif. (2538 kc coast, 2142 kc ship) and the Los Angeles-San Diego, Calif. (2598 kc coast and 2206 kc ship) areas.** The Commission also modified a score of ship radiotelephone station licenses by deleting the frequency 2738 kc because of interference caused from second harmonic radiation to aeronautical frequency 5476.5 kc. Also, the applicability of provisions of Sect. 351(a)(1) of the Communications Act imposing new radio installation and operator requirements on certain ships between 500 and 1600 gross tons, was deferred until Jan. 1, 1955 by the FCC.

Technical

The "beginning of a revolution in battlefield communications" is how Maj. Gen. George I. Back, Army Chief Signal Officer, described the Army's newest tactical weapon—combat television—at a demonstration at Ft. Meade, Md. The Interim Tactical TV System, developed by the Radio Corp. of America under a contract with the Signal Corps, used three hand-held television cameras operating out of ¾ ton trucks. The signal from these cameras was microwaved back to a base station mounted in a 2½ ton truck, which contained the control and monitoring equipment for the system. This base station is designed to be located at a Regimental or Division Command Post. An L-20 DeHavilland airplane mounting two TV cameras and a

(Continued on page 84A)

* The data on which these NOTES are based were selected by permission from *Industry Reports*, issues of August 9, 16, 23, 30 and September 6, published by the Radio-Electronics-Television Manufacturers Association, whose helpfulness is gratefully acknowledged.

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The facts on delivery today are as follows: We're swamped with orders. We can only handle your minimum requirements. Larger orders will be filled as promptly as General Electric's greatly-expanded production lines swing into "high". So place your order promptly. A shipment of G-E junction transistors applied in *your circuits* will save space and power, and reduce weight . . . as they deliver the important design advantages listed at the right.

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WELDED SEAM CONSTRUCTION...free from solder-flux contamination.

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HERMETIC SEAL...unaffected by moisture.

HIGH TEMPERATURE OPERATION...rated for a maximum junction temperature of 100°C.

LONG LIFE...designed for long-term, stable performance.

SMALL SIZE...extremely compact design provides added flexibility for all applications.

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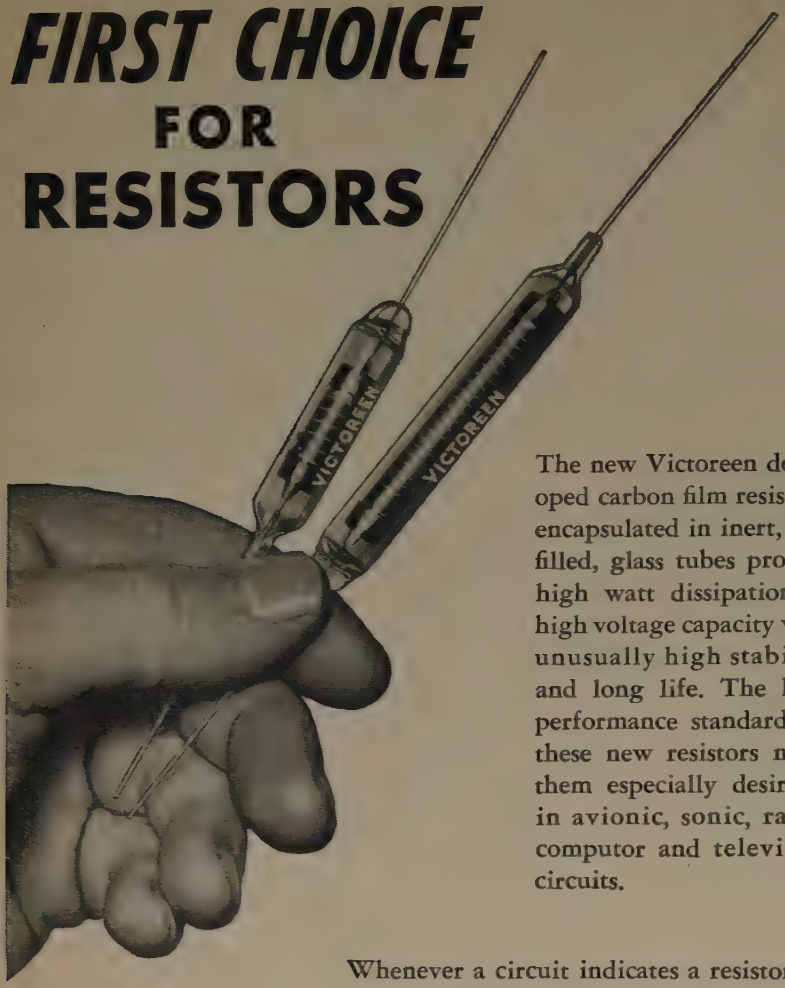

The new Victoreen developed carbon film resistors, encapsulated in inert, gas-filled, glass tubes provide high watt dissipation or high voltage capacity with unusually high stability and long life. The high performance standards of these new resistors make them especially desirable in avionic, sonic, radar, computor and television circuits.

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Whenever a circuit indicates a resistor for extraordinary high performance standards, specify Victoreen as your first choice for stability and long-life. From engineering, thru all manufacturing processes these resistors are made to the highest possible quality. To many manufacturers, the use of Victoreen resistors has meant the elimination of costly field failures.

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Industrial Engineering Notes

(Continued from page 82A)

transmitter formed a part of the system with its signals received by the base station. The electronics industry was represented at the demonstration by personnel from General Electric Co., Motorola Inc., Sylvania Electronic Products, Inc., Philco Corp., AT&T and Bell Laboratories, Inc., in addition to RCA. . . . **The Radio Technical Commission for Aeronautics has released four new reports of studies on airborne radio equipment.** Three of the reports recommend minimum performance standards for, respectively, radio marker receivers operating on 75 mc, ILS glide slope receivers, and ILS localizer receivers. They are part of a series of studies conducted by Special Committee 58 relating to criteria for radio equipment intended for use in aircraft operating under instrument flight rules. Each report contains an appendix outlining procedures for testing equipment to determine its performance. The fourth report, by SC60, concerns VOR test signals. The following reports are available through RTCA, Building T-5, Washington 25, D. C.: VOR Test Signals. A description of methods of determining, in an aircraft, the accuracy of VOR bearing indications, together with an analysis of the causes of VOR bearing error due to VOR receiver malfunctioning. Paper 66-54/DO-56, 50 cents. Minimum Performance Standards—Airborne Radio Marker Receiving Equipment Operating on 75 mc. Paper 87-54/DO-57, 30 cents. Minimum Performance Standards—Airborne ILS Glide Slope Receiving Equipment. Paper 88-54/DO-58, 30 cents. Minimum Performance Standards—Airborne ILS Localizer Receiving Equipment. Paper 89-54/DO-59, 30 cents. . . . **Since size and weight reduction of electronic equipment is becoming increasingly important in many applications, particularly in military equipment, the National Bureau of Standards, through its continuing program for the development of miniaturization techniques and their application to airborne electronic equipment, has published a circular, "Miniature Intermediate-Frequency Amplifiers."** The circular thus makes information on one phase of the program, the miniaturization of radar components, readily available to the electronics industry. Three miniature high-gain, high-frequency, intermediate-frequency amplifiers (20 to 100 mc) were developed with particular emphasis on the use of circuit elements suitable for maximum design simplicity, circuit flexibility and ease of manufacture. The units are about one-eighth the size and one-half the weight of the equipments they supersede, the Bureau announced. The circuitry was designed with emphasis on the use of subminiature tubes and their application in low-noise input circuits. Some units are hermetically sealed for protection against contamination and moisture, and to provide for operation under the extreme temperature range (−65 to 200 degrees centigrade) required for

(Continued on page 86A)

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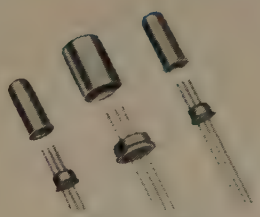

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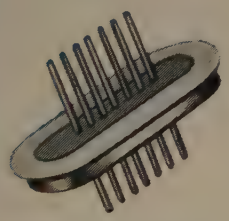
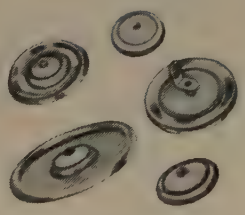


HERMETICALLY SEALED TERMINAL WITH CUSHIONED GLASS CONSTRUCTION



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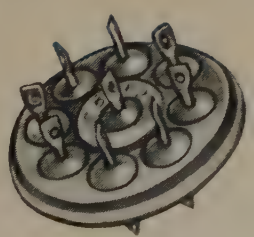

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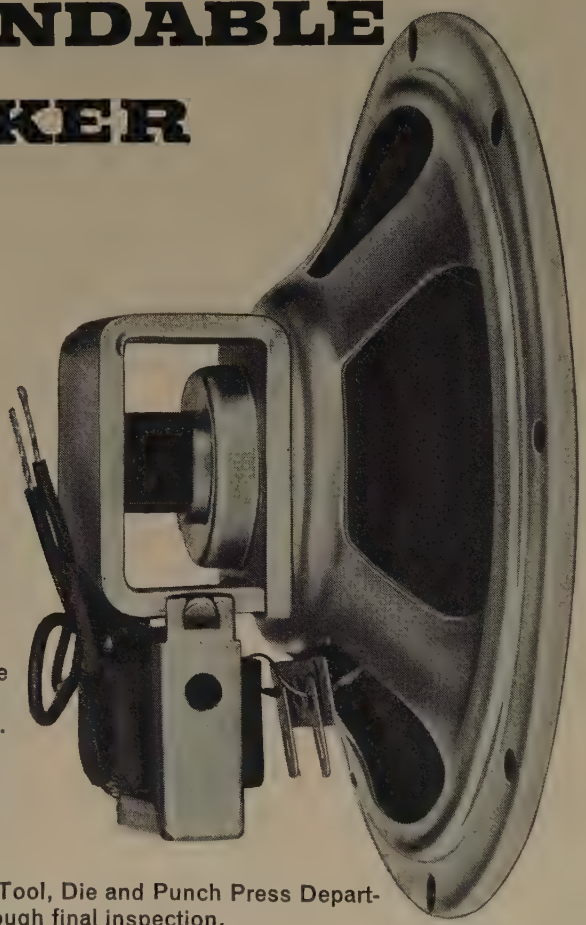

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Industrial Engineering Notes

(Continued from page 84A)

such equipment. "Miniature Intermediate-Frequency Amplifiers" is available at 40 cents per copy through the Superintendent of Documents, Government Printing Office, Washington 25, D. C.

Television

By the end of this year, nearly 100 cities will have television network facilities capable of handling color signals the American Telephone & Telegraph Co. reported to the Federal Communications Commission. In a letter advising the Commission of the progress being made in this field, AT&T reported that there were 47 cities which had been equipped for network color service. By the end of the year, AT&T said, its present plans call for installation of necessary equipment in an additional 46 cities. . . . **An extensive investigation of the radio-television industry loomed recently when the Senate Interstate and Foreign Commerce Committee approved the hiring of two persons to conduct the investigation.** With this move, the plan of the Communications Subcommittee to appoint an ad hoc committee to study the recommendations made during the recent UHF hearings died aborning. The two investigators, according to the announcement, will "make a study of possible federal regulations of radio and TV networks" as proposed in S. 3456, introduced earlier by Sen. John Bricker (R., Ohio), Chairman of the Senate Interstate and Foreign Commerce Committee. However, the study will consider all means of creating a truly, nation-wide competitive system of television, including the factors played by the availability of UHF receivers and transmitting equipment. . . . **According to FCC figures, 595 construction permits were issued for TV stations between the lifting of the freeze and the end of August.** Of these, 93 have been deleted, leaving 502 post-freeze CPs outstanding on Sept. 1. Of the CPs issued, 32 were for educational stations and seven of these have received special temporary authority to start operating. In addition to these seven educational outlets, FCC has granted STAs to 299 commercial stations, 164 on VHF channels and 135 using UHF channels. . . . **With the increase in the amount of color programming, the American Telephone & Telegraph Co. announced that it now can transmit color programs to 65 cities across the nation.** An AT&T spokesman said that a large number of facility installations were completed in the last few weeks, raising the total by a wide margin from the 47 cities reportedly interconnected for color on Aug. 1. In the 65 cities, AT&T reported that 95 of the 145 stations are receiving color in addition to black and white programs. By the end of the year, AT&T estimated that 95 cities could be equipped with network transmission facilities to handle color and that

(Continued on page 88A)

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strengthen our defenses

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HHB 3358 PRESSURIZED RECEPTACLE

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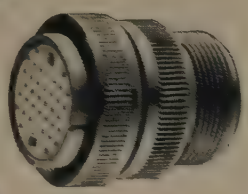

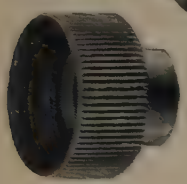

HHB 3310 TUBING CLAMP

—for limited space requirements

This compact connector assembly (approx. 2⅜-in. long x 1¾-in. dia.) is a forerunner of miniature AN Components design to be pioneered by HHB design and engineering departments.

Components include 38 contact receptacle and plug—two 35 Amp. (No. 12 wire) and thirty-six 5 Amp. (No. 18 wire) contacts. Shell adapter and clamp of gray anodized aluminum. Features the new HHB 3300 Series flexible plastic tubing clamp instead of standard AN cable clamp.

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Industrial Engineering Notes

(Continued from page 86A)

125 stations would be receiving this service. . . . **The Federal Communications Commission was asked by the Fairmont Broadcasting Co. and WJPB-TV, Fairmont, W. Va., to take jurisdiction over all community television distribution systems operating for a profit.** The petition charged that community TV systems across the nation are now operating in direct competition with authorized TV outlets, and without control or restraint of federal or state government. The firm said that "many community TV systems are acting as communications common carriers for the purpose of providing off-the-air pickup and relay of TV broadcast signals from distant TV stations to subscribers for profit . . . and many are inserting commercials in network programs without the consent of the stations from which the programs come. "In other areas," the petition continued, "community TV distribution systems are preparing to telecast live, local advertiser-sponsored programs over their closed circuits, which will constitute a serious economic threat to allocated and established TV stations in the areas served by these systems." The request asked the FCC to define the existence or extent of its juris-

(Continued on page 90A)

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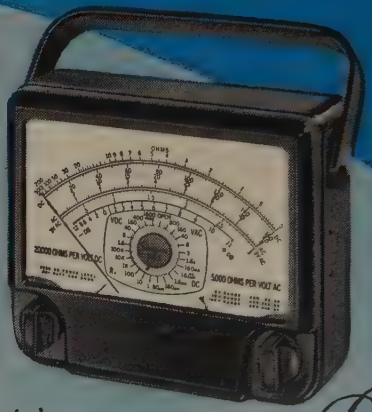
the new vom with a 7″ meter



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20,000 Ohms per volt DC. 5,000 Ohms per volt AC. 33 RANGES

DC VOLTAGE: 0-1.6, 0-8, 0-40, 0-160, 0-400, 0-1600, 0-4000 volts (20,000 ohms per volt sensitivity)
AC VOLTAGE: 0-3, 0-8, 0-40, 0-160, 0-800 volts (5,000 ohms per volt sensitivity)
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VOLUME LEVEL IN DECIBELS: −12 to +45.5 DB in 4 ranges. Zero DB Power Level, .001 watt in 600 ohms.
DC RESISTANCE: 0-500 ohms (4.5 ohms center); 0-5,000 ohms (45 ohms center); 0-50,000 ohms (450 ohms center); 0-500,000 ohms (4,500 ohms center); 0-5 megohms (45,000 ohms center); 0-50 megohms (450,000 ohms center)
DC CURRENT: 0-80, 0-160 microamperes, 0-1.6, 0-16, 0-160 milliamperes, 0-1.6, 0-16 amperes (267 millivolts maximum drop)
MODEL 262 complete with 2 test leads with removable alligator clips, 4,000 v. DC multiplier
Dealer's Net Price, including Adjust-A-Vue Handle..**$59.50** Carrying Case..**$9.95**
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Industrial Engineering Notes

(Continued from page 88A)

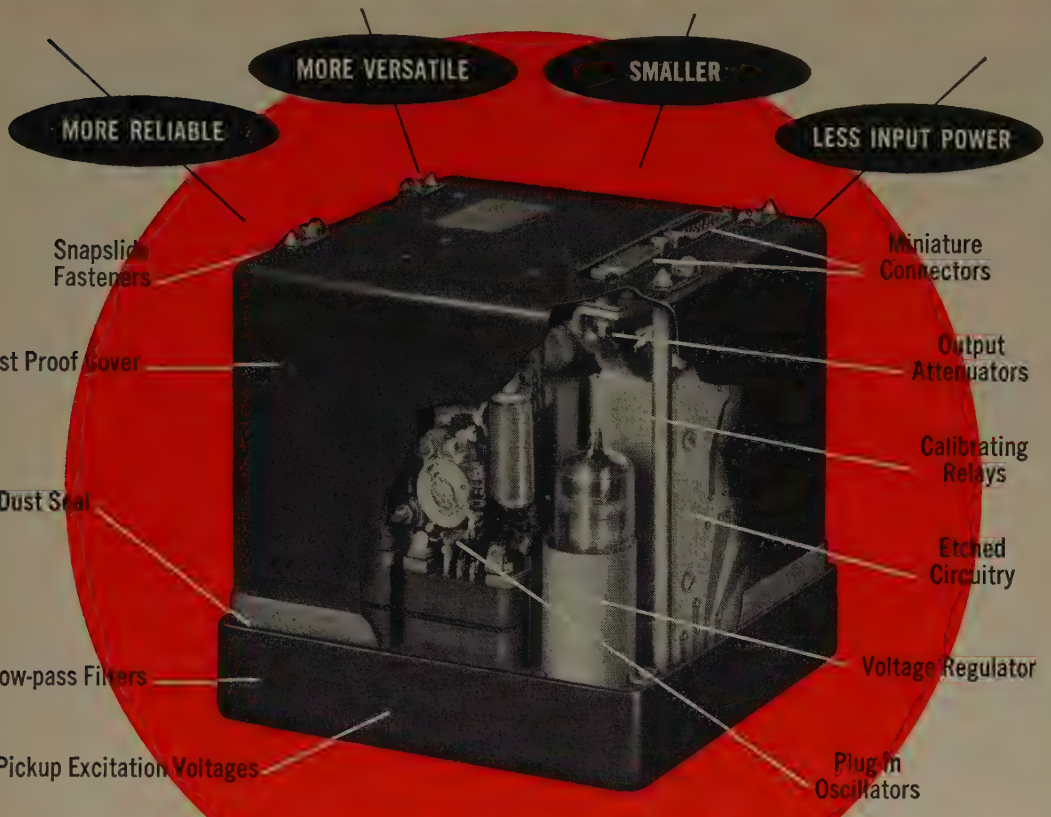
diction with respect to the installation and operation of any such TV system, to the end that the same rules and orders of the Commission, applicable to TV stations, shall apply to community TV distribution systems. . . . **As the telephone company announced plans to join the largest number of cities for any month this year to its television networks in September,** FCC records showed that nearly 600 television construction permits have been issued since the freeze. There also are 189 applications pending FCC action. At the same time, the Commission has granted permission to approximately 300 TV outlets to start operation since the lifting of the freeze. Including the 108 pre-freeze stations, total authorizations made by the FCC reached 703 on Sept. 1. Even with 93 proposed outlets deleted, 610 stations held TV permits at the beginning of this month.

RETMA Activities

Additional copies of the report from the Conference on Reliability of Electrical Connections still are available. The conference was held April 15–16 at the Illinois Institute of Technology under the sponsorship of the RETMA Engineering Department with the cooperation of the Aircraft Industries Association. The printed report of the conference, which was devoted to methods of connecting to the terminations of wires and cables to produce equipment free of the difficulties caused by loose or broken connections, includes not only the papers presented, but also transcripts of question and answer periods. Several supplementary articles relating to connection reliability also are included. The report of the Conference on Reliability of Electrical Connections is priced at $5 per copy and may be ordered through the RETMA Engineering Department, 500 Fifth Ave., New York 36, N. Y. . . . **RETMA has filed a petition with the Federal Communications Commission requesting additional time to permit completion of technical studies of radiation from restricted radiation devices before the Commission acts on a proposal to amend Part 15 of the FCC rules governing such devices.** Announced also was an order of the Commission, which had been approved, extending the date for filing comments until Oct. 15. RETMA, and a number of the other 25-odd petitioners in the rule making proceeding, requested a longer extension of time to complete various technical studies relating to the complex proposals covered in Docket No. 9288. An FCC spokesman told RETMA that the two-month extension was a routine action covering the entire Docket and that more time could be granted where additional study is needed on specific sections. . . . **Dates for the Symposium on Printed Wiring and Circuitry have been changed to Jan. 20 and 21, Virgil M. Graham, Associate Director of the RETMA Engineering Department,**

(Continued on page 92A)

A NEW TELEMETER STANDARD



BENDIX-PACIFIC TATP-4&5 PACKAGES

These compact Bendix-Pacific Telemetering Packages offer users of telemetering systems a **better** means to instrument such quantities as pressure, force, temperature, voltage, acceleration and vibration.

The units are smaller and various combinations may be used to provide compact multi-channel systems of up to 18 subcarrier bands. Each unit operates on **unregulated** +150 VDC and 28 VDC since it contains its own voltage regulating circuits. Each unit may be provided with individual relays for switching oscillator inputs from signal to calibrate position. Model TATP-4 contains four and Model TATP-5 contains six separate and independent regulated +5 VDC excitation voltage sources for use with resistance type pickups.

No vacuum tubes are required in mixing the outputs of the individual subcarrier oscillators to the composite audio signal for direct modulation of the RF transmitter, thus greatly improving the reliability of the system at a point where a tube failure would affect all subcarrier channels.

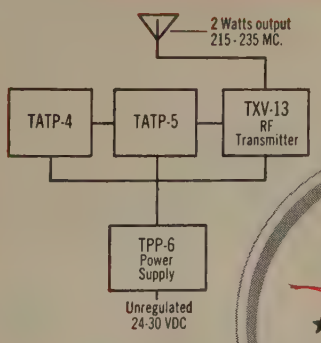
Standard Bendix Model TPP-6 Power Supplies are available to handle from one to three of these telemeter packages plus a 2 watt crystal controlled RF transmitter. Many types of interchangeable subcarrier oscillators are also available. RF amplifiers are available for increased power outputs up to 100 watts.

CHARACTERISTICS

	TATP-4	TATP-5
No. of bands	4	6
Subcarrier bands	1.7 kc. to 70 kc.	1.7 kc. to 70.0 kc.
Oscillator Types	TOE-30V, TOE-31V, TOR-8V, TOR-9V, TOL-9V	Same as TATP-4
Input Voltages	6, 12, or 24 VDC ±10% 150 VDC $\pm\frac{10}{5}$%	6, 12, or 24 VDC ±10% 150 VDC $\pm\frac{10}{5}$%
Weight (Less Oscillators)	Approx. 2 lbs.	Approx. 3 lbs.
Dimensions	5.0" x 5.0" x 4.5"	5.0" x 7.5" x 4.5"

Write for Complete Information

TYPICAL 10 BAND SYSTEM



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SYSTEM STABILITY

SERVO MOTORS: Servo motors with high torque to inertia characteristics possessing (built-in) inherent damping ranging in size from ¾" to 1¾" diameter are available. Low speed, low power motors for use in simple instrument servos where high damping and/or low time constant is required can also be provided.

VISCOUS DAMPED SERVO MOTORS: Provide integral viscous damping for simple instrument servos. Any degree of damping can be provided. These units reduce no load speed of standard motors to 50% or 75% of normal, providing 70% or 50% of critical damping respectively.

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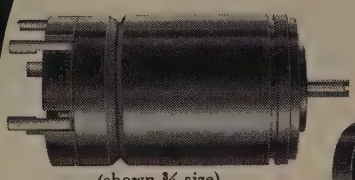
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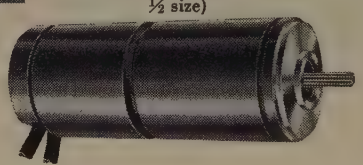

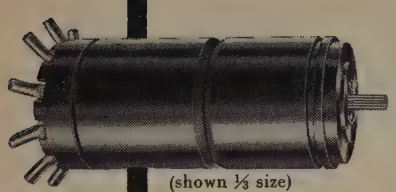
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INERTIAL DAMPED MOTORS: Integral inertially damped motors for use in high speed and/or high gain servo systems—damping on acceleration or deceleration basis with little loss in normal no load speed. These units make possible system cut off frequencies up to 25 cps using magnetic amplifiers.

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Industrial Engineering Notes

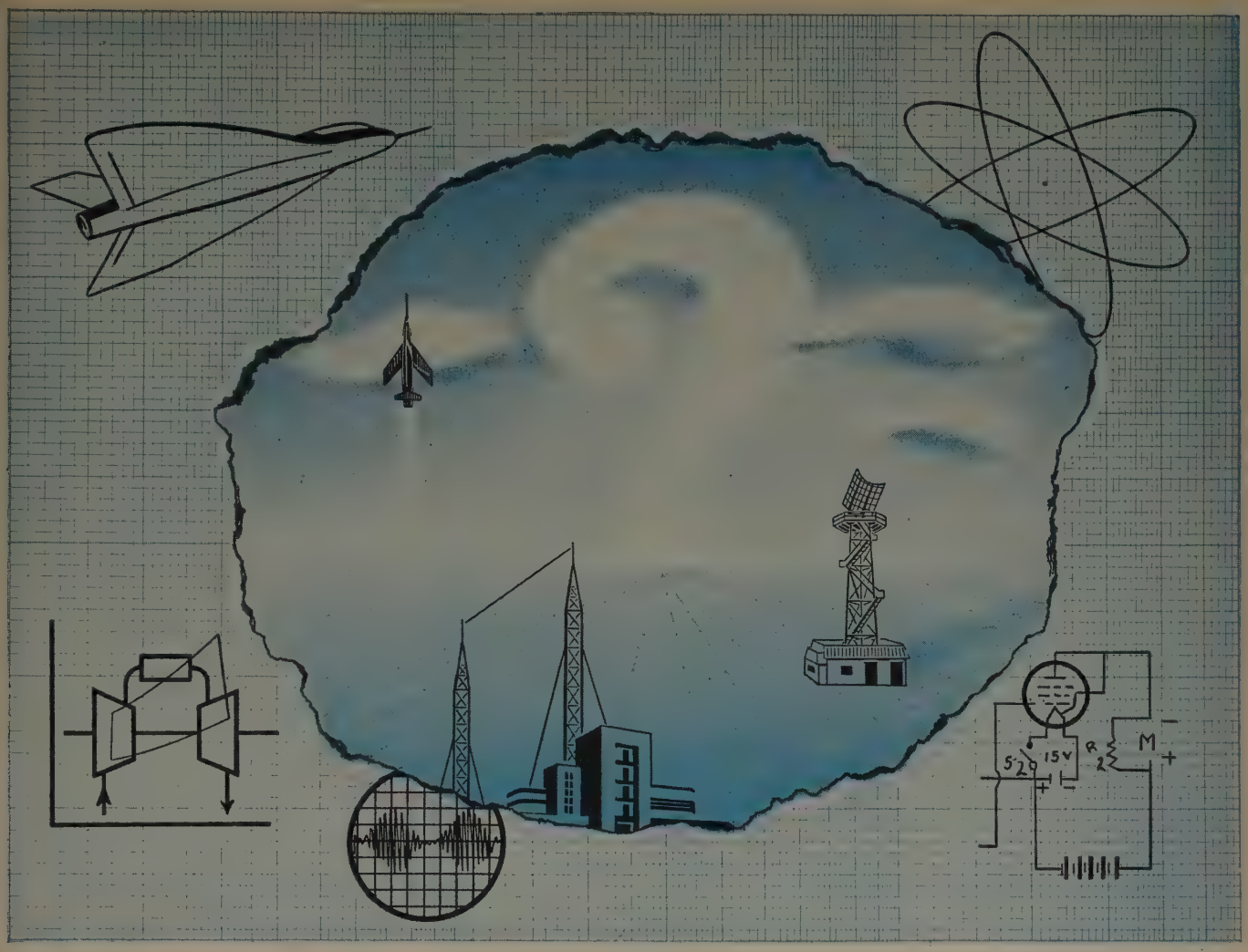
(Continued from page 90A)

announced. The symposium will be held at the University of Pennsylvania, Philadelphia, under sponsorship of the RETMA Engineering Department. Mr. Graham also announced that Donald W. Cottle of the General Electric Co., Electronics Park, Syracuse, N. Y., should receive all technical papers which are proposed to be presented at the symposium. . . . **With the view of promoting a more thorough understanding of the practical techniques, problems and applications of printed circuits and to provide an opportunity for discussion of recent developments, the RETMA Engineering Department will sponsor a Symposium on Printed Circuits, Jan., 20–21, at the University of Pennsylvania, Philadelphia.** The technical program for the two-day meeting is under the direction of Donald W. Cottle, General Electric Co., Syracuse, N. Y. A broad outline of the technical program, including six sessions, has been announced in tentative form by Mr. Cottle as follows: I. Product Design Applications. This session will include papers on the conversion of specific products from conventional construction to printed circuit construction. Problems of conversion and superiority or shortcomings of printed products as compared to conventional apparatus will be covered. Techniques of design of printed circuit products also will be the subject of papers. II. Reliability and Serviceability. This session will feature papers describing techniques of meeting Underwriters Laboratories' requirements and specific reliability data and special tests for reliability. Papers concerning repairmen's acceptance of printed circuit products and their problems and repair techniques also will be given. III. Management Considerations. This session will include papers on converting a manufacturing operation from conventional electronic products to printed circuit products. It also will contain papers on economics. IV. Panel Session on Techniques of Producing Printed Wiring Boards. This panel will consist of a Chairman and one member to speak as an advocate of each system of producing printed wiring panels (etched, stamped, fired-on paint, etc.). V. Printed Components and Components for Use With Printed Wiring. This session will feature papers on materials, printed components and components for use with printed wiring. VI. Production Techniques and Manufacturing Methods. Papers describing quality control and inspection, assembly equipment, soldering techniques, etc., will be offered during this session. Papers presented at the Symposium will be published in an official Proceedings which will be made available to the public following the two-day technical meeting.

AERONAUTICAL

The Civil Aeronautics Administration and the Air Navigation Development Board have joined in an experiment to

(Continued on page 94A)



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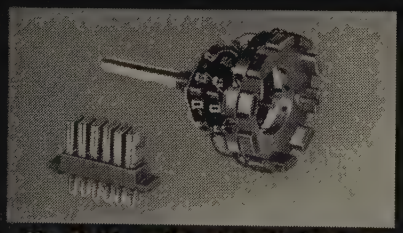
New M-21 "McMite" Sub-Miniature Quartz Crystal

with Pins

Latest addition to the McCoy line of high quality, precision quartz crystals is the new M-21 "McMite" . . . with .040" dia. pins. The new M-21 supplements its twin brother, the M-20 "McMite" which has wire leads. Together, these two crystals offer extreme adaptability to designers of multi-channel communications and frequency control equipment. The M-21 plugs into a sub-miniature socket while the M-20 can be wired or soldered to a printed circuit terminal board. Both may be incorporated into miniature selector switch assemblies. The M-20 and M-21 crystals are sub-miniature, hermetically sealed units which deliver the same performance as larger, regular size crystals, *yet require only one-fifth as much space.* Both meet military specification requirements for fundamental operation above 5 mc. and overtone operation above 15 mc. Neither sacrifices stability or dependability but meets the same characteristics and performance requirements as the conventional larger crystal units.



Six M-21 units plugged into a 1¾" strip of sockets (left) and ten M-20 units wired into a 2" dia. switch assembly (right) illustrate the space saving qualities of the McCoy "McMite" crystals.

McCoy ELECTRONICS CO.

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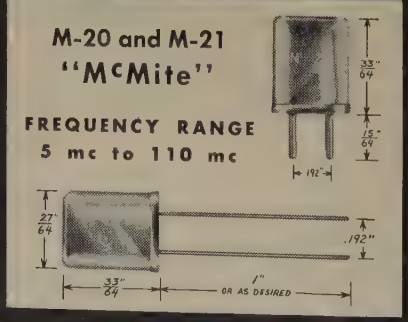
M-20 and M-21 "McMite"

FREQUENCY RANGE 5 mc to 110 mc



Industrial Engineering Notes

(Continued from page 92A)

determine if increased efficiency and coverage can be obtained by the use of a 200-foot tower to house a VOR (very-high frequency omni-directional radio range), F. B. Lee, CAA Administrator, announced. The proposed VOR installation will be located at Harrellsville, N. C., a heavily wooded area, and will be constructed and tested by the CAA's Technical Development and Evaluation Center at Indianapolis. If the experiment proves successful, the CAA Administrator said, the Harrellsville facility will be commissioned and other high-tower VOR's installed where siting or distance coverage is a problem.

Electronic Parts

Military representatives of the three services to the Advisory Group on Electron Tubes are: G. R. Kilgore (Army), Signal Corps Electronic Laboratory; Robert Grantham (Navy), Bureau of Ships, and Amos Dicke (Air Force), Wright Air Development Command. The complete civilian roster of the six working groups is as follows: *Microwave Tubes*—Wellesley Dodds, RCA Laboratories; H. William Welsh, Jr., University of Michigan, and W. C. Brown, Raytheon Manufacturing Co. *Receiving Tubes*—J. O. McNally, Bell Telephone Labs., and Everett Coon, Tung-Sol Electric Inc. *Tube Techniques*—R. J. Boniley, Research Laboratory, General Electric Co., and R. E. Palmateer, Sylvania Electric Products Inc. *Special Tubes*—Dr. Sidney T. Smith, Electron Tube Laboratory, Hughes Aircraft Co. *Power and Gas Tubes*—Stuart T. Martin, WCAX Broadcasting Corp., Burlington, Vt. *Semiconductor Devices*—Dr. Harvey Brooks, Physics Department, Harvard University; A. E. Anderson, Bell Telephone Labs., and Dr. Harper Q. North, Pacific Semiconductor Inc. Under a charter drawn up for the Electronic Parts group, but not yet issued, provisions have been made for establishing working groups. It was made known that the fields to be covered by these groups, none of which have been organized or staffed, will consist of the following: Automation Techniques, Capacitors, Coils-Inductors-Transformers, Electric and Magnetic Materials, Electro-Mechanical Devices, Frequency Control Devices, Transmission Lines, and Resistors. These working groups will, according to the charter, consider the impact of military requirements on R&D activities and take appropriate action to encourage adoption of joint requirements where military needs are similar. Furthermore, they will evaluate R&D activities for adequacy of timing in view of strategic guidance, technical limitations and advances, and potential threats to national security. Additionally, measures to improve reliability of all electronic parts within each respective field of interest will be considered by the groups.

(Continued on page 96A)

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"Precision airborne communication equipment must have a dynamotor of consistently reliable quality . . . that's why we specify Winco."

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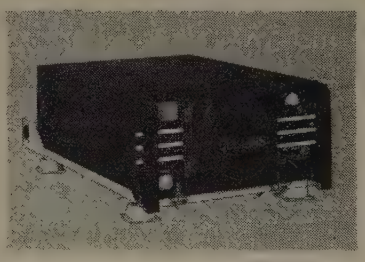
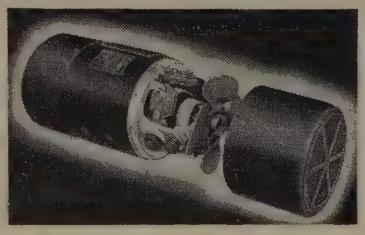


Collins 18S Transmitter/Receiver, used for reliable HF communications in major air-lines and executive aircraft uses a Winco-Engineered Frame 51 Dynamotor.



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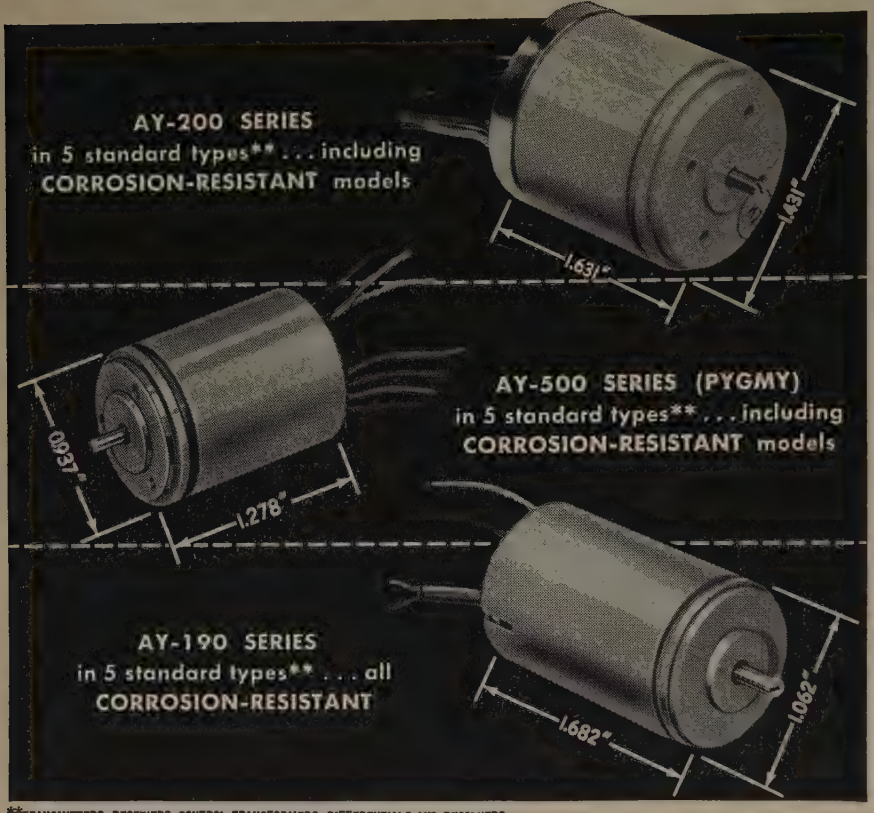

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Industrial Engineering Notes

(Continued from page 94A)

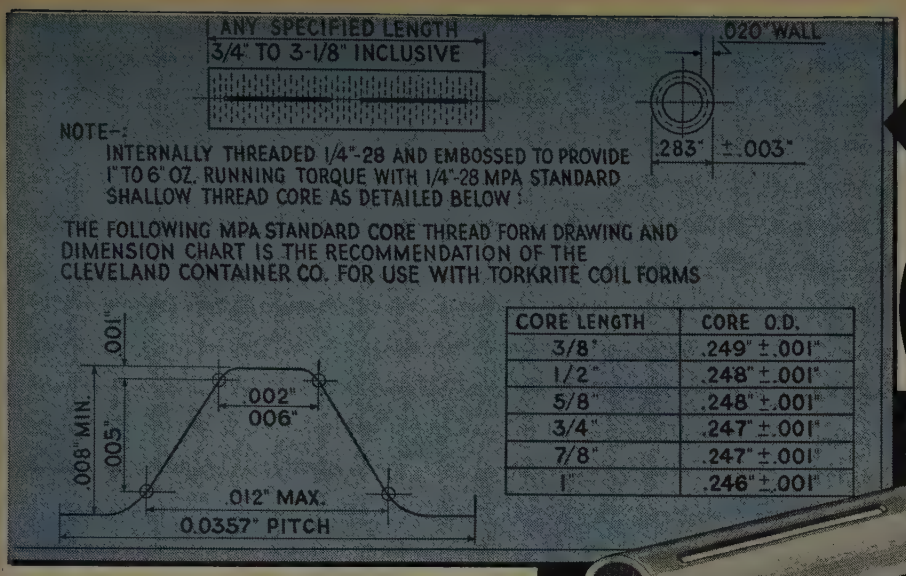
MATERIALS

Out of 12 materials considered by the Department of Defense to deserve the highest priority of conservation attention, eight are used extensively by the electronics industry. The long-range outlook for most of these materials places them in a position of continuing short supply and in the case of selenium and tantalum a more aggressive conservation effort by the Defense Department is urged. The highest priority of conservation has been ordered within the defense establishment for the following materials used by the industry: Nickel, molybdenum, tantalum, selenium, copper, cobalt, tungsten and columbium.

AUDIO

The Cincinnati Chapter of the Professional Group on Audio held five meetings at the Engineering Society Headquarters, with E. J. Jones, Chairman. On November 17, 1953, a tapescript, "How Much Distortion Can You Hear?," was presented by E. M. Jones, W. M. Gulden, and J. P. Goode, the Executive Committee of the Chapter. Demonstration of pure and distorted tones was made to check the perception of various amounts of harmonic and IM distortion. Frequencies of 100 cps and 1 kc were used, and the audience kept score sheets to indicate their choices between pure and distorted tones. The tapescript was forwarded to the PGA Tapescript Committee. On January 19, 1954, a paper, "All Transistor Hearing-Aid" was presented by S. K. Webster of Beltone Corp. Mr. Webster discussed the problems in design and manufacture of a hearing-aid using transistors to replace vacuum tubes completely. The advantages of transistors in hearing-aids were pointed out, especially as regards battery requirements and life. On March 16, 1954, J. P. Goode, Manager, National Sound Service spoke on "Electronic Church Bells." Mr. Goode discussed tones produced by cast bells and means of reproducing these tones electronically, especially by recordings. There was a demonstration of bell tones produced in various ways and recordings of tuned chimes music. On April 20, 1954, the paper was "Wow and the Measurement Thereof." The speaker, M. Young, Assistant Sales Manager, AC Motors Div., General Industries, discussed the causes and cures of wow. Several eccentric turn tables, producing various percentages of wow, were demonstrated to indicate the major cause of wow. Both 1-kc tones and music were used in the demonstration. On

(Continued on page 98A)


ANY SPECIFIED LENGTH
3/4" TO 3-1/8" INCLUSIVE
.020" WALL
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NOTE—:
INTERNALLY THREADED 1/4"-28 AND EMBOSSED TO PROVIDE 1" TO 6" OZ. RUNNING TORQUE WITH 1/4"-28 MPA STANDARD SHALLOW THREAD CORE AS DETAILED BELOW :
THE FOLLOWING MPA STANDARD CORE THREAD FORM DRAWING AND DIMENSION CHART IS THE RECOMMENDATION OF THE CLEVELAND CONTAINER CO. FOR USE WITH TORKRITE COIL FORMS
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.005"
.002"
.006"
.012" MAX.
0.0357" PITCH
CORE LENGTH | CORE O.D.
3/8" | .249" ± .001"
1/2" | .248" ± .001"
5/8" | .248" ± .001"
3/4" | .247" ± .001"
7/8" | .247" ± .001"
1" | .246" ± .001"

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- permits use of lower torque since it is completely independent of stripping pressure.
- recycling ability is unmatched.
- is stronger mechanically because of heavier wall.
- provides 1-6" oz. running torque when used with MPA standard shallow thread core.
- has no holes or perforations thru tube wall which eliminates cement leakage locking cores.
- has smooth adjustment of core without lubricant.
- torque increases less after winding as heavier wall reduces any tendency to collapse and bind core.
- maximum stability results as core cannot move in relation to winding after peaking as it is engaged in internal threads.
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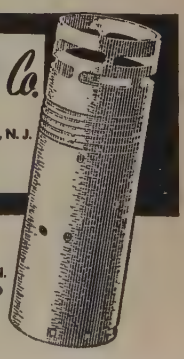

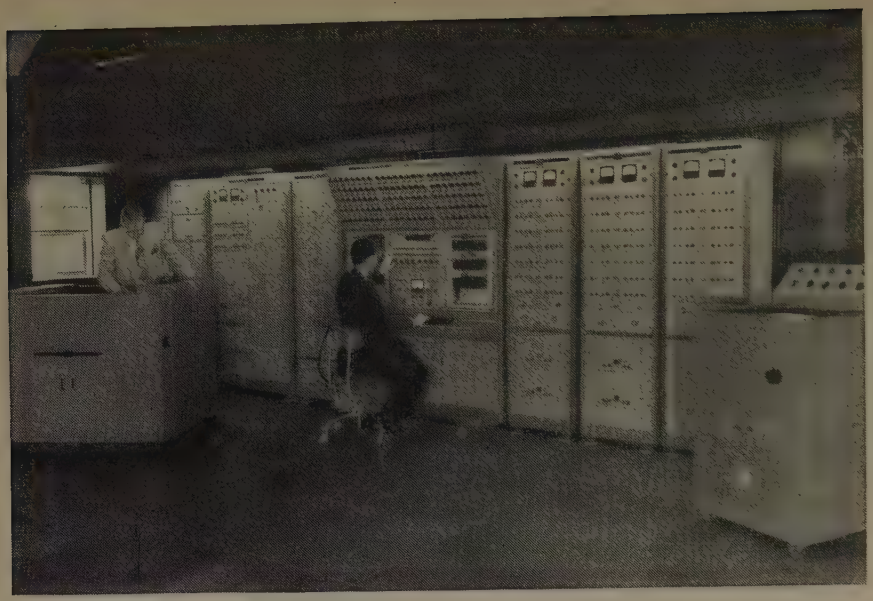

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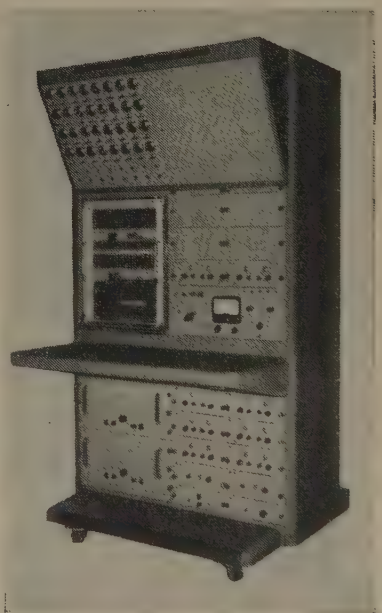

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PROFESSIONAL GROUP MEETINGS

(Continued from page 96A)

June 20, 1954, at Carillon Park, Dayton, Ohio, a field trip was made to hear the Carillon and "Celestron." The "Celestron" is a high-power (1,200w) audio system housed in the carillon tower at Dayton. A high-fidelity system, it broadcasts music throughout the park, using both disc and tape. A half-hour concert on the cast bell carillon was followed by a concert on the "Celestron." The members enjoyed picnic lunch before the program and afterwards inspected the "Celestron" equipment. The Albuquerque-Los Alamos Chapter met on August 23 at the Radiation Therapy Building in Albuquerque; D. V. Couden, the chairman, presided. Ed. May, of the International Electronics Company in Dallas, spoke on "Frazer-May Speaker Designs." Discussion followed a demonstration of the latest speaker designs.

Nuclear Science

On June 8 at Mitchell Hall, University of New Mexico, the Albuquerque Chapter met. Barney J. Carr, Sandia Corporation, spoke on the properties of ores containing radioactive material and the instruments used in prospecting these ores. He also presented the operating theory of Geiger-Mueller tubes and scintillating crystals. A discussion of electronic circuits and components used in these instruments followed the talk. The Los Alamos Chapter convened on August 10 at P-Division Conference room for a business meeting. The following committee chairmen were appointed: membership, Walter H. Weber; meetings and papers, Richard J. Watts and Thomas J. Boyd; publicity and arrangements, Richard Hiebert. On August 17, the Chapter again met under the chairmanship of William Briscoe. Kenneth Urbach, of Dumont Laboratories, spoke on "The Mono-Accelerator Tube," and Bernard Linden, also of Dumont, spoke on "Photomultiplier Construction."

Electronic Computers

The New York-Long Island Chapter has held five meetings this year. At their first meeting, on January 26, C. Andrews, E. L. Schmidt, and C. Ammond delivered a paper on "Reservisor." At the next meeting W. D. Groth spoke on "The AMA System." On April 27 the Chapter met at the Bell Telephone Laboratories in Murray Hill, New Jersey, and C. Andrews, the Vice-Chairman, presided. The papers presented were "Diad" and "Ferro Electrics" by W. A. Malthaner and J. R. Anderson. On May 24 A. A. Katz, D. E. Keefer, and P. H. Rosenthal spoke on "The Univac System." At the June meeting new officers were elected for 1955 and W. W. Wockenfuss spoke on "High Speed Printing."

IRE People

Ira Kamen (M'48–SM'52) has been named vice president in charge of engineering, research, and development for the Brach Manufacturing Corporation, Division of the General Bronze Corporation. Formerly vice president in charge of sales, he will continue with this work in addition to his new responsibilities.

I. KAMEN

From 1948 to 1950 Mr. Kamen was manager of the TV department for Commercial Radio Sound Corporation, an RCA distributor of engineering products. While there he developed the RCA Antenna-plex outlet system. Previous to 1948 he was consultant to a number of organizations we well as an office of the Intravideo Corporation, where he directed development of the first TV master antenna system.

During the war years, Mr. Kamen was a supervisory professional engineer for the Third Naval District, directing the installation of radio, radar, and sonar systems on naval combat and transport vessels. He was cited by the Bureau of Ships and the Commandant, Navy Yard, New York, for having performed outstanding service. Over 100 technical articles and several books have been written by Mr. Kamen.

❖

Emmett R. Shute (M'25–SM'43) has retired as Vice-President of The Western Union Telegraph Company to become President of the Serial Federal Savings and Loan Association of New York City. Manhattan's oldest savings and loan association, it serves 21,000 members and has resources of more than $23,000,000.

Joining Western Union as assistant to the Traffic Engineer, New York, in 1912, he pioneered many of the company's outstanding technical advances. He was early to recognize the role of engineers as business executives, and four whom he trained are now vice presidents of the telegraph company. During World War II, Mr. Shute was chief liaison officer between Western Union and the Armed Services with the rank of Lieutenant Colonel and was Chairman of the Telegraph Committee of the Board of War Communications. He was awarded a Chief Signal Officer's Certificate of Appreciation.

Mr. Shute received the B.S. degree from Purdue University. He is a Fellow of the American Institute of Electrical Engineers, and Vice President of the New York Chapter, Armed Forces Communications Association. He has served as president and director of a number of Western Union's subsidiary companies, and director of Ramp Buildings Corporation.

(Continued on page 102A)

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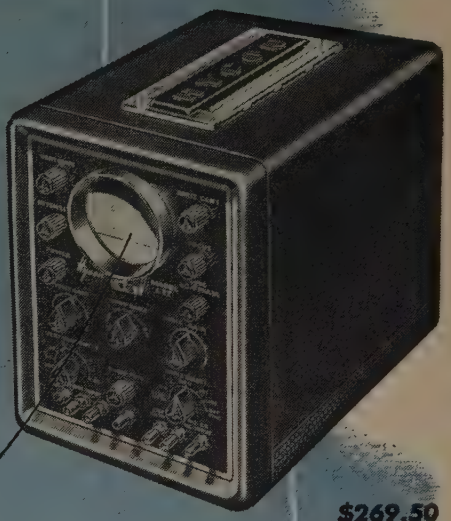
MODEL 617



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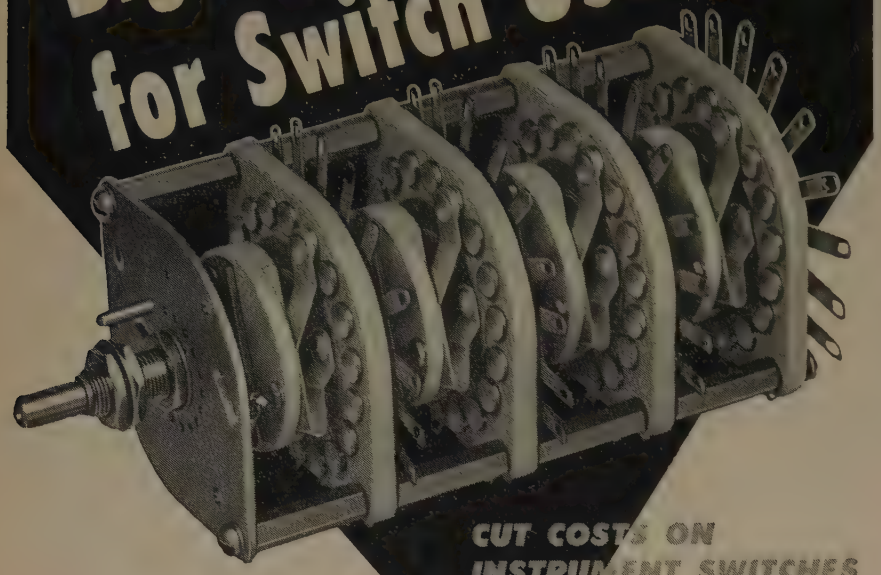

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IRE People

(Continued from page 101A)

Emil A. Sellars (A'47), former chief engineer of Greenville television station WGVL, has been named to the same position at WBTW, the Jefferson Standard Broadcasting Company's new top power TV station in Florence.

A native South Carolinian, Sellars was active in the installation of the Greenville station and has directed its engineering facilities since it went on the air. Before moving to Greenville, he was assistant supervisor in the video control operation of another Jefferson Standard station, WBTV, Charlotte, N. C. While serving with WBTV, Sellars was video-engineer for the first TV mobile pickup in the Carolinas, the telecast of the 1951 Shrine Bowl football game in Charlotte. Sellars has had numerous specialized assignments in the electronics field including work with Pan American Airways in the Canal Zone in 1941 and preparation of anti-submarine defenses at the Norfolk Naval Air Station in 1942. He helped promote and initiate night radar ship-bombing against the Japanese and installed and operated an air-sea rescue communications station in the Los Angeles area.

Sellars is a graduate of U. S. Navy radio-radar schools and of Capitol Radio Engineering Institute in Washington, D. C., where he was trained in the theory and technical aspects of television. He later served as an instructor at the Institute.

❖

Leroy Packer (S'48–A'51) will direct the new Computer Division of the Polarad Electronics Corporation. This division will develop and apply electronic computer techniques, services and equipment for industrial and government use.

L. Packer

Mr. Packer was previously a research engineer at Franklin Institute Research Laboratories, where he did extensive work on airplane simulators and developed a new device for the electronic testing of abrasive wheels. At Burroughs Adding Machine Company, 1950–1952, he did research on high speed input devices for large scale digital computers, and did transistor circuit development. Mr. Packer then joined the Columbia University Electronics Research Laboratory. Here, he directed research on navigational and guidance systems, and developed digital devices for use in the tracking of military aircraft.

He received the B.S. degree in e.e. from Columbia University and the M.S. degree from the Moore School of the University of Pennsylvania. Several of his papers have been published in technical journals. He is a member of Tau Beta Pi.

(Continued on page 105A)

IRE People——

(Continued from page 102A)

Albert C. Gable (SM'46), who has been appointed manager of product planning for the new marketing section of the General Electric Industrial and Transmitting Sub-Department.

A. C. Gable

He joined G.E. at Schenectady in 1929 in the Test Department, and a year later transferred to the company's vacuum tube operations. He held varied engineering assignments in industrial tube development before heading the industrial and transmitting tube engineering staff in 1948.

Mr. Gable is a native of Macon, Ga., and was graduated from Georgia Tech with the B.S. degree in E.E. In 1952, he received the first Service Award of the industry's Joint Electron Tube Engineering Council, following two years' service as chairman of the council's critical materials committee.

Four staff members of the Diamond Ordnance Fuze Laboratories recently have received Certificates of Achievement from Lt. Col. J. A. Ulrich, the commanding officer, and Wilbur S. Hinman, Jr., the technical director of the laboratories. The Diamond Ordnance Fuze Laboratories is a newly activated ordnance research and development installation. Its primary mission is development of fuzes and fuzing systems for the Department of the Army.

Henry P. Kalmus (A'19-SM'45) was awarded a Certificate of Achievement for "the invention of basic new electronic systems in electronic ordnance and for the invention of a new system for measuring liquid and gas flow." Born and educated in Vienna, he was head of the Orion Radio Laboratories until 1939. After two years with the Emerson Radio Corporation, Mr. Kalmus was associated with the Zenith Radio Corporation where he developed the Cobra phonograph pickup and a novel lightmeter based on magnetic modulation of photocells.

Paul E. Landis (A'44-SM'53), an electronic scientist at the laboratories, was presented a Certificate of Achievement "for liaison with foreign scientific missions resulting in more effective prosecution of development programs."

A Certificate "for the outstanding development of a new method of field instrumentation" was presented to **Glenn L. Scillian** (SM'53). Mr. Scillian was associated with a commercial concern before joining the staff of the Ordnance Development Division, National Bureau of Standards, which in 1953 became the Diamond Ordnance Fuze Laboratories.

(Continued on page 106A)

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for use with multi-conductor cables

These new Bendix*-Scinflex waterproof plugs are a modification of our standard AN type "E" (environment resistant) connector. They are designed to meet all "E" performance requirements when used with multi-conductor cables. Each plug includes a modified AN3057B cable clamp which provides inward radial compression on multi-conductor cables. This unique feature completely eliminates cable strain—a common source of circuit trouble.

In addition, there are gaskets at all mating surfaces and an accessory sleeve is available to accommodate an extreme range of cable sizes. A folder describing this new waterproof plug—and the various sizes in which it is manufactured—may be obtained by writing our Sales Department.

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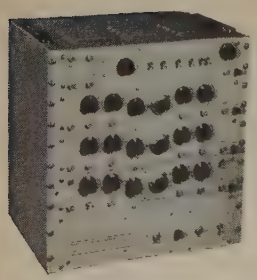

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Telephones: EXbrook 8-6764 and TExas 0-8006

IRE People

(Continued from page 105A)

Thomas A. Prugh (A'46–M'52) was presented a Certificate of Achievement "for the outstanding application of Transistors in circuitry." Mr. Prugh is an electronic scientist at the laboratories.

❖

Dr. J. E. Hobson, director of Stanford Research Institute, has announced the appointment of **Dr. Donald L. Benedict** (SM'53) as director of physical sciences research. He was formerly assistant director of the engineering division. He will supervise all project work and Institute sponsored research in the fields of chemistry, chemical engineering, metallurgy, ceramics, biochemistry and physics.

D. L. Benedict

With SRI since 1949, Dr. Benedict has supervised programs in electronics research and has participated in theoretical and technical work in the application of new materials to electronic components. He has served as consultant to Raytheon Manufacturing Co., advising on microwave tubes, design of equipment for microwave cooking and dielectric heating.

In 1945 he was named a research fellow in Harvard University's Department of Engineering Science and Applied Physics. Earlier he was a research physicist with Sylvania Electric Products Research Laboratory at Flushing, L. I. At the University of Wisconsin he received the Ph.D. in physics, specializing in experimental nuclear physics; he remained at the university to work on atomic energy development. He is a graduate of Knox College, Galesburg, Ill.

Dr. Benedict is president of SRI's branch of RESA (Scientific Research Society of America). He is a member of American Physical Society, American Mathematical Society, and Institute of Aeronautical Sciences.

❖

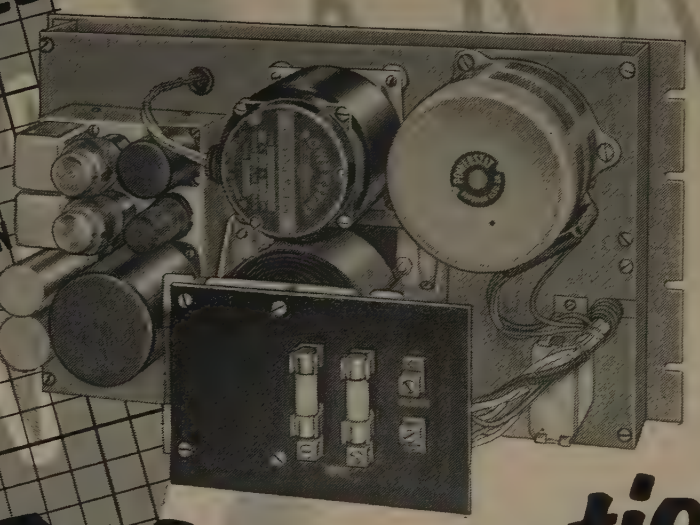
Dr. J. E. Hobson (M'45) director of Stanford Research Institute, has formally succeeded to the national presidency of Eta Kappa Nu, electrical engineering honor society with some 22,000 members throughout the world. He will preside at Eta Kappa Nu's assembled convention at the time of its Golden Anniversary observance at the University of Illinois on October 15 and 16.

The anniversary convention will be attended by the national executive council and national advisory board, delegates from 56 college chapters, undergraduate groups, delegations from alumni chapters, past presidents, and other officials. The

(Continued on page 108A)

Maintain Constant Output Voltage

THE WAVE THAT GOES IN



COMES OUT UNCHANGED

STABILINE
Type EM4102R

with Zero Waveform Distortion

If you want to maintain constant output voltage with zero waveform distortion or constant voltage to large industrial loads, you will find the STABILINE Type EM the answer.

STABILINE automatic voltage regulators Type EM not only stabilize and regulate the voltage output and exactly reproduce the input waveform, but feature:

- **outstanding efficiency — comparable to the most conservatively designed fixed-ratio transformers**
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Manufacturers of: Powerstat Variable Transformers • Stabiline Automatic Voltage Regulators • Voltbox A-C Power Supplies • Powerstat Light Dimming Equipment • Varicell D-C Power Supplies • Superior 5-Way Binding Posts

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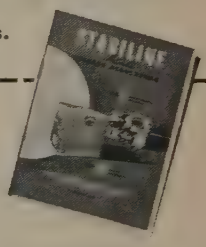

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Please send Bulletin S351 on STABILINE automatic voltage regulators.

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Company Address.....

City.....*Zone*.....*State*.....

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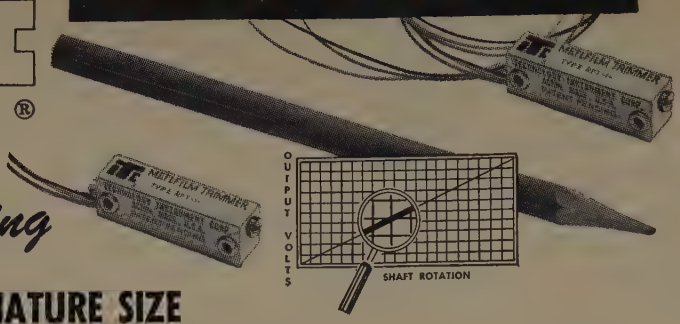

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MINIATURE SIZE

INFINITE RESOLUTION plus "ZERO PHASE-SHIFT"

DIMINUTIVE SIZE

. . . (approximately ⅜" square end surface), permits stacking seven units in a square inch of panel area. Ideal for trimming adjustments in computers, analyzers, telemeter and airborne electronic equipment.

Rugged construction insures dependability despite wide changes in ambient temperature and extreme conditions of salt spray, humidity and vibration.

INFINITE RESOLUTION

. . . of the unique deposited metal resistance element embodied in Type RFT Metlfilm Trimmer Potentiometer is available over a wide resistance range. 9000° of adjustment, the equivalent of 25 turns of the adjustment screw on which sliding contact rides, permits voltage settings to be set and maintained with extreme precision.

SPECIFICATIONS

ELECTRICAL

RESISTANCE RANGE: 50-25,000 ohms
TOTAL RESISTANCE TOLERANCE: ±10%
INDEPENDENT LINEARITY: ±5% of total resistance
RESOLUTION: Infinite
POWER RATING: ½ watt at 40°C., ¼ watt at 125°C. per JAN-R-19 test specification.
AMBIENT TEMPERATURE RANGE: −65°C to +125°C.
TEMPERATURE COEFFICIENT OF RESISTANCE ELEMENT: .000250/°C (nominal)
DIELECTRIC TEST: 500 volts DC between all leads, shaft and mounting eyelets for 5 seconds without flashover or breakdown.

MECHANICAL

RESISTANCE ELEMENT: Metal film deposited on inert base.
MECHANICAL ROTATION: 26 complete turns (nominal).
USABLE MECHANICAL ROTATION: 90% minimum of slider travel is on resistance element.
END STOPS: Will withstand 1 inch pound maximum applied torque.
VIBRATION: Exceeds exacting requirements of MIL-E-5272a.

For further details write:

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535 Main St., Acton, Mass. COlonial 3-7711
West Coast Mail Address P.O. Box 3941 North Hollywood, Calif. POplar 5-8620
Plant, 11020 Sherman Way, Sun Valley, California

IRE People

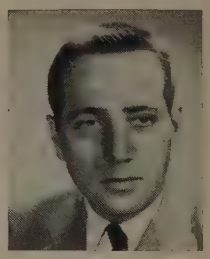
(Continued from page 106A)

new president has issued special invitations to the four living members who founded Eta Kappa Nu at Urbana in 1904, M. K. Akers, C. K. Brydges, W. T. Burnett, and E. B. Wheeler.

Dr. Hobson became a member of Eta Kappa Nu while an undergraduate at Purdue University. He obtained his doctorate at California Institute of Technology, was director of the Department of Electrical Engineering at the University of Pittsburgh and professor of electrical engineering at Illinois Institute of Technology. He was director of Armour Research Foundation, Chicago, for four years before assuming leadership of Stanford Research Institute in March 1948.

❖

Dr. Leonard C. Maier (M'52) of Syracuse has been appointed manager of engineering for the General Electric Cathode Ray Tube Sub-Department. In his new position he will be in charge of all cathode ray tube product engineering for G-E monochrome TV picture tubes, color TV picture tubes, and special purpose cathode ray tubes.



L. C. Maier

Following a year as a research physicist, Dr. Maier joined G.E. in 1950 as an engineer in the Electronics Laboratory. He served as assistant section engineer for communication systems and then as supervisor for multiplex systems before being assigned to the company's Heavy Military Electronic Equipment Department as supervisor for digital techniques and computers. Earlier this year, he was appointed manager of applications and techniques for the laboratory's development engineering section, a position he held until his present appointment.

He served two years as a Navy electronics officer, and upon his return to civilian life in 1946 entered Massachusetts Institute of Technology for graduate study. He took his master's degree there in 1948 and his doctorate in 1949, majoring in electromagnetic and nuclear theory. While at MIT he served as a research associate in the institute's Research Laboratory for Electronics. A native of LaGrange, Ill., Dr. Maier was graduated from Phillips Exeter Academy, Exeter, N. H., and then from Williams College in 1944.

Dr. Maier is a member of the American Physical Society, the Research Society of America, Sigma Xi, Theta Delta Chi, and Phi Beta Kappa.

(Continued on page 112A)

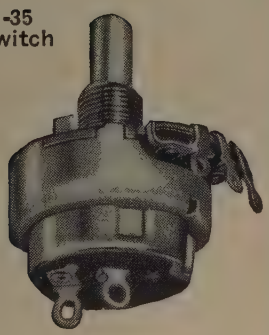
RESISTORS...FOR EVERY NEED

Meets JAN-R-94 type RV3

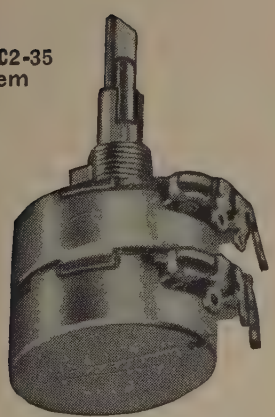
1/2 watt 1-1/8" diameter variable composition resistor. Also available with other special military features not covered by JAN-R-94.

TYPE 35

TYPE GC-35 With Switch



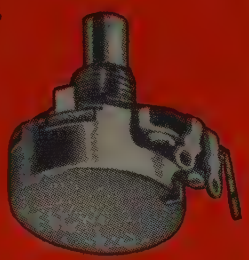
TYPE C2-35 Tandem



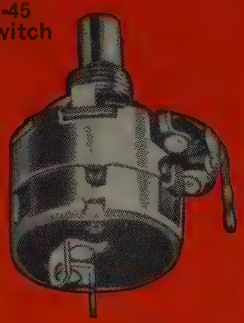
Meets JAN-R-94 type RV3

1/4 watt 15/16" diameter variable composition resistor. Also available with other special military features not covered by JAN-R-94.

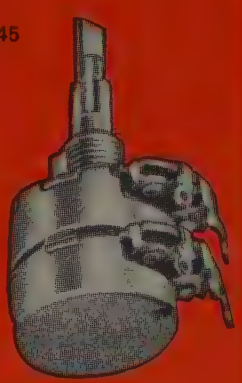
TYPE 45



TYPE GC-45 With Switch



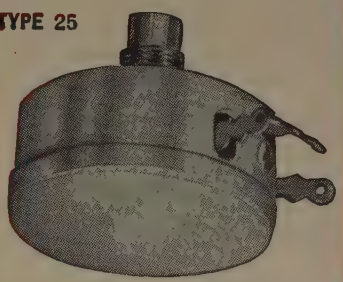
TYPE C2-45 Tandem



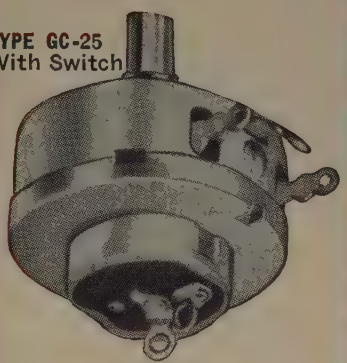
Meets JAN-R-19 types RA25 and RA30

4 watt 1-17/32" diameter variable wirewound resistor. Also available with other special military features not covered by JAN-R-19.

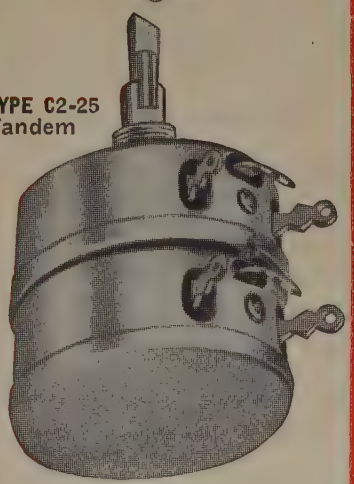
TYPE 25



TYPE GC-25 With Switch

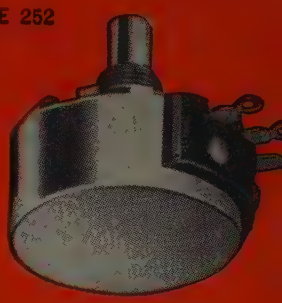


TYPE C2-25 Tandem

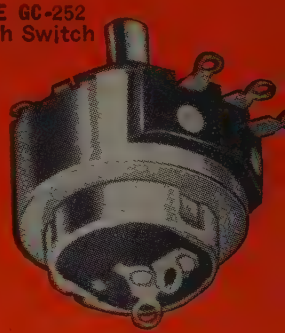


2 watt 1-17/64" diameter variable wirewound resistor. Also available with other special military features not covered by JAN-R-19.

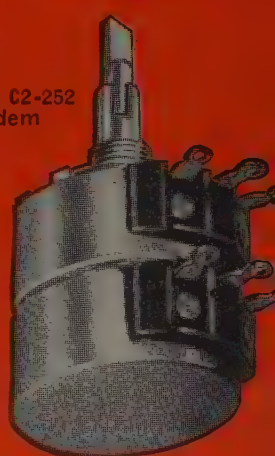
TYPE 252

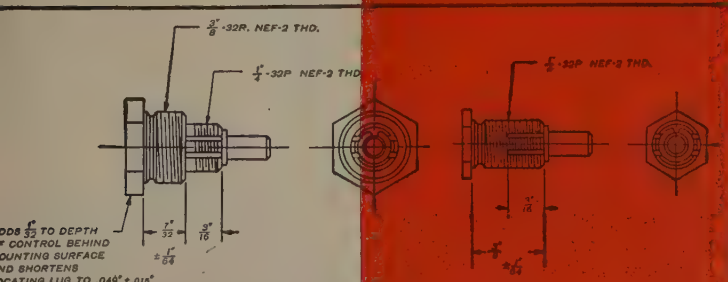


TYPE GC-252 With Switch

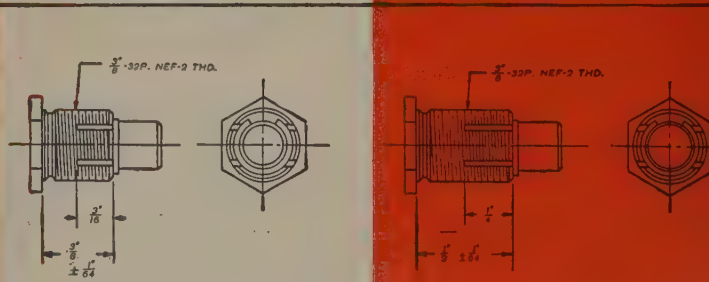


TYPE C2-252 Tandem



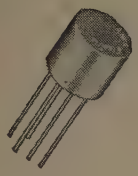


LOCKING BUSHINGS FOR TYPE 65 CONTROL.



LOCKING BUSHINGS FOR CONTROL TYPES 25, 252, 95, 35, 90, 45.

Specialists in Precision Mass Production of Variable Resistors • *Founded 1[illegible]*



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PULSE TRANSFORMERS

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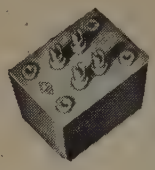

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TRANSFORMERS

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DEPOSITED CARBON RESISTORS

After intensive study of the many special requirements arising in modern resistor applications, Welwyn has succeeded in developing a resistor-encapsulating process employing a new, insulating material. When thus applied, this material will withstand an applied potential of 8000 volts. Its insulating effectiveness is in excess of 100,000 megohms.

Welwyn insulated resistors can be placed contiguous to each other without the need for additional insulating material. In all other respects, these new insulated resistors offer the same performance and quality advantages generally associated with Welwyn Deposited Carbon resistors. Available in ¼, ½, 1 and 2 watt ratings.

Manufactured in England and Canada

For complete data and specifications write to Dept. KL-4

WELWYN
High Stability
DEPOSITED CARBON RESISTORS

ROCKBAR CORPORATION 215 East 37th Street, New York 16, N. Y.

IRE People

(Continued from page 108A)

Thomas B. Friedman (A'46–M'46–SM'50), television engineer and consultant, is joining the Adler Communications Laboratories of New Rochelle, New York. He will assist in the development and application of satellite television stations and television station systems in which fields Adler Laboratories has a rapidly expanding program.

T. B. FRIEDMAN

Mr. Friedman is leaving the Empire Coil Company of New Rochelle where he was Chief Television Engineer and instrumental in the planning and construction of WXFL, Cleveland; KPTV, Portland: and KCTY, Kansas City.

Mr. Friedman is a graduate of the University of Michigan, with the B.S. degree in e.e. He is a member of the American Institute of Electrical Engineers, and the Society of Motion Picture and Television Engineers and an Associate Member of the Association of Federal Communications Consulting Engineers. He is also a consulting engineer and a Registered Professional Engineer in the State of Ohio.

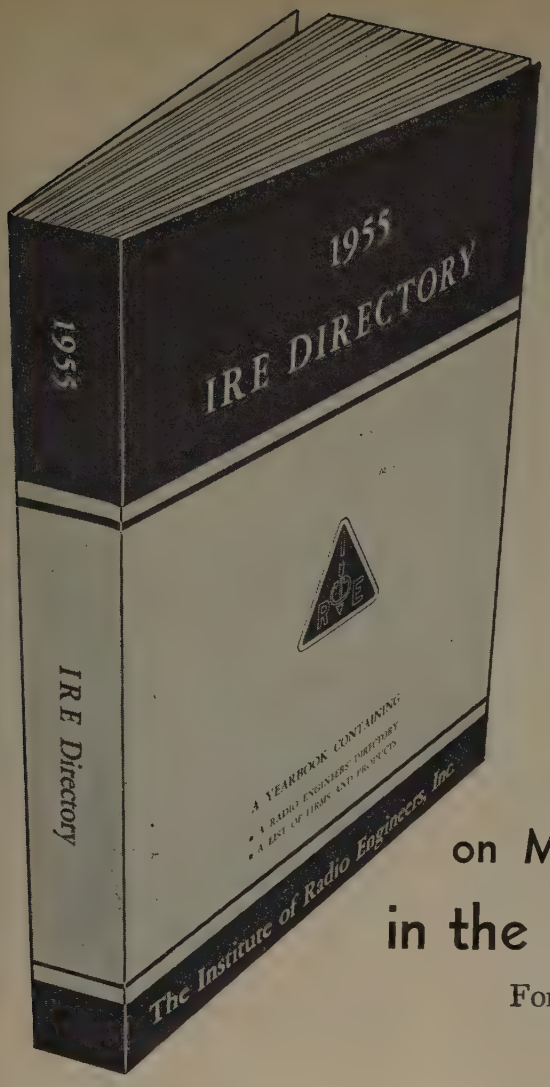
❖

Victor H. Fraenckel (A'36), former liaison scientist in physics at the General Electric Research Laboratory, has been appointed consultant on scientific relations, Dr. C. Guy Suits, vice president and director of research, announced. Mr. Fraenckel will be responsible for informing and counseling the management of the research laboratory on scientific work outside the General Electric Company. He will evaluate and forecast the probable future trends in external scientific activities of significance to the company.

Prior to joining G.E. in October 1937, he was engaged in electronics research for Farnsworth Television Co., in Philadelphia, and as a physicist for HyGrade-Sylvania in Emporium, Pa. During World War II he was appointed an expert consultant to the Secretary of War and served in Europe on special missions concerned with electronic counter-measures and scientific intelligence. For this work, he received several citations and was awarded the Medal for Freedom, a special civilian award of the War Department.

Mr. Fraenckel is a native of Chicago, attended Hyde Park High School in that city, and the Lycée Condorcet and the École Polytechnique in Paris, France before completing graduate work at the Universities of Chicago and Michigan.

(Continued on page 114A)



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tubeless

ELECTRONIC VOLTAGE STANDARDS

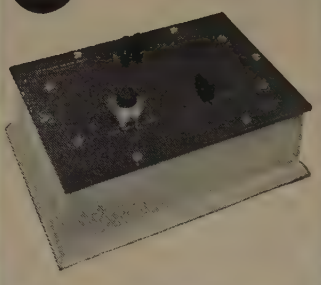
For calibration of meters and oscilloscopes—*Model VRS-AC10*

This new instrument can be used for direct checking and calibrating of vacuum tube voltmeters. It can also be used for indirect checking of power meters.

Range: 1 mv to 100 mv
100 mv to 1 v
1 v to 10 v
Accuracy: ±0.25%
Input: 115 VAC ±10%, 1ø, 60∿
Output loading: ½ megohm or higher. However, not damaged by short circuit.

We solicit inquiries for AC standards of other voltage and current ratings.

AC



DC voltage standard
Model VRS15

The primary application of Sorensen's new tubeless DC standard is as a component in recording equipment, where it eliminates a power supply and a standard cell. It is also intended for use as a laboratory standard, illustrated below.

Output: 1½ VDC at 6 ma. Not damaged by short circuit.
Output impedance: 12 ohms.
Accuracy: ±1 mv against line
Ripple: 350 microvolts
Stability: a 24-hour stability check shows output drift well under 1 mv
Input range: 95-130 VAC, 1ø, 60∿

Other DC standards can be designed to your requirements up to 5 VDC and other currents. Special mechanical configurations of the DC standard will be designed for your particular applications.

Write for complete information.

Model VRS-AC10 and Model VRS15 are the latest additions to Sorensen's standard line of power regulating equipment.

Send for your free copy of the new general catalog — write to Sorensen & Co., Inc., 375 Fairfield Ave., Stamford, Conn. In Europe: Sorensen A.G., Gartenstrasse 26, Zurich 2, Switzerland.

DC

SORENSEN

IRE People

(Continued from page 112A)

Albert A. Pulley (A'25–VA'39), manager, general recording, Victor Records Division, RCA has been elected president of the Audio Engineering Society. He succeeds Jerry B. Minter, president of Components Corp., Denville, N. J.

A. A. PULLEY

The new president has been with RCA and its predecessor company since 1920, when he joined the test department of General Electric. He transferred to the radio department, working first on radio transmitters and then on the GE disc recording system. When RCA assumed these interests in 1930, Mr. Pulley shifted to Camden, N. J., where he did design engineering work on electrical-mechanical devices, including disc recording systems. He subsequently became manager of recording and was promoted to his present post this year. In 1949, Mr. Pulley received the RCA Award of Merit for his part in the development of 45 rpm recording.

A member of the Society since its founding, Mr. Pulley became a fellow in 1952 and was elected executive vice president a year ago. He will preside as general chairman at the organization's Sixth Annual Meeting, to be held in the Hotel New Yorker, New York, October 14–16.

The following transfers and admissions were approved to be effective as of October 1, 1954:

Transfer to Senior Member

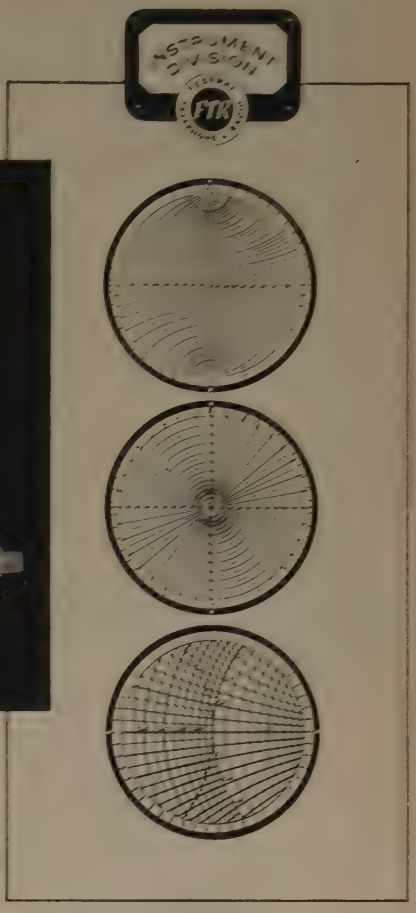
Anderson, T. J., 8133 Pleasant Plains Rd., Towson 4, Md.
Arnold, J. B., 220 Cheryl Dr., San Antonio 1, Tex.
Barcala, J. M., San Martin 379, Buenos Aires, Argentina
Breitwieser, G. F., Jr., 37 E. Clinton St., Clayton, N. J.
Burke, J. J., 5022 Ladd Ave., Los Angeles 32, Calif.
Crain, C. M., Engineering Bldg. 149, University of Texas, Austin, Tex.
DeLany, R. H., 5000 Euclid Ave., Cleveland 5, Ohio
Egerton, W. G., Box 1161, San Antonio 6, Tex.
Ing, G. W., 135 Ostrom Dr., San Antonio 12, Tex.
Jackson, K. R., 12443 Deerbrook Lane, Los Angeles 49, Calif.
Muschamp, R. A., 103 Sandra Lane, N. Syracuse, N. Y.
Nash, H. E., Box 258, Quaker Hill, Conn.

(Continued on page 116A)

Federal introduces the FIRST instrument for

PLOTTING IMPEDANCE instantaneously!

Z-g DIAGRAPH



Eliminates complex, time-consuming calculations... all measurements read from interchangeable charts

The Z-g Diagraph, by Rhode and Schwarz—one of West Germany's leading instrument makers—is probably the most unusual device ever developed for fast, high-accuracy impedance and admittance measurements.

The Diagraph indicates results directly on a Smith, or similar, chart. It can obtain an entire broad-band characteristic in a few minutes . . . present it visually . . . without tedious calculating.

Separate, quickly-interchangeable charts permit measuring of reflection coefficient, resistance and reactance (Smith Chart), or magnitude and phase angle of an unknown impedance. Either can be used without recalibration or change in test set-up.

The transmission characteristic of a four-terminal network, such as a filter or attenuator, can also be read directly. The latter feature permits the Diagraph to be used for measuring the phase angle between two voltages of unequal amplitude but of the same frequency.

For further information on the performance and versatility of the Z-g Diagraph, mail the coupon to Federal's Instrument Division.

SPECIFICATIONS:

Frequency Range: Type ZDU . . . 30 to 300 mc.; Type ZDD . . . 300 to 2400 mc.

Characteristic Impedance: 50 ohms.

Measuring Range: Impedance . . . 1 to 2500 ohms; Phase . . . 0 to 360°; Attenuation . . . 0 to 30 db.

Accuracy: Amplitude . . . ±3%; Phase . . . ±1.5°.

Terminals: Type N.

Power Supply: 115 volts (or 220 volts), 50 to 60 cycles.

Weight: 135 lbs.

Price: $5,950 net, FOB, Clifton, N. J.

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Federal Telephone and Radio Company
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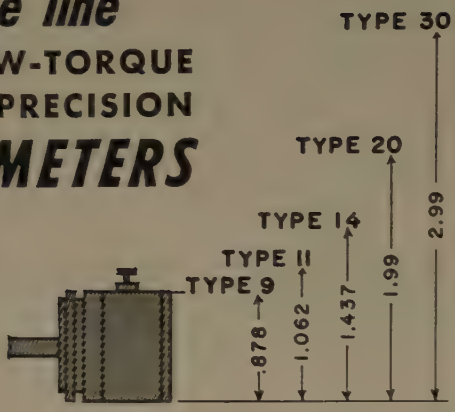

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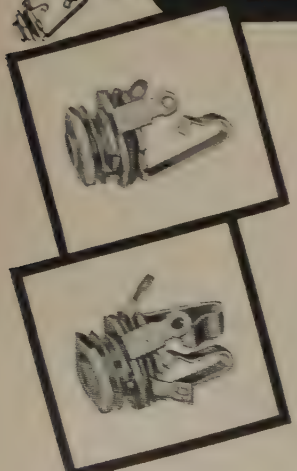

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Rives, J. B., 1721 Kentucky Ave., San Antonio 1, Tex.
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Stevens, F., Jr., 1525 Addison Rd., Palos Verdes Estates, Calif.
Symonds, R. J., 1860 Berkeley Ave., Pomona, Calif.
Tarrodaychik, P., 215 Christine Dr., San Antonio 10, Tex.
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Tull, E. H., 19 McDonald Ave., London, Ont., Canada

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DeDiemar, J. L., 3316 Jewell St., San Diego 9, Calif.
Gindsberg, J., 275 Ft. Washington Ave., New York 32, N. Y.
Higa, W. H., Jet Propulsion Laboratory, 4800 Oak Grove Dr., Pasadena, Calif.
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Koga, I., 254, 8-Chome, Kami-Meguro, Tokyo, Japan
Lakhanpal, D. D., 29 New Queens Rd., Bombay 4, India
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Grimmett, C. A., 113 Iroquois Lane, Liverpool, N. Y.
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Miller, C. E., 320 N. Ash St., Wichita 7, Kans.

(Continued on page 119A)

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Maring, K. T., 133 Old Port Isabel Rd., Brownsville, Tex.
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Spratley, J. B., Ellerson, Va.
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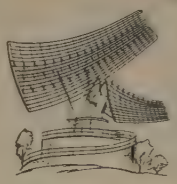

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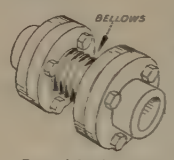

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(Continued from page 119A)

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Anderson, A. F., 93 Whitney Road, Medford 55, Mass.
Aubel, L. J., 172-11—83 St., Jamaica 32, L. I., N. Y.
Avers, B., 1605 S. Main St., Los Angeles 15, Calif.
Barber, W. S., 1138 W. Eighth St., Santa Ana, Calif.
Benningfield, L. M., 1734 N. Yale, Wichita, Kan.
Bentley, D. E., 832 Cree Avenue, Akron 5, Ohio
Blanchard, W. A., 29 Glenview Dr., West Orange, N. J.
Brooks, C. L., Jr., 221 Linden Ave., Wood Lynne 6, N. Y.
Brown, H. R., 603 Crescent Dr., Bound Brook, N. J.
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Dawson, R. W., Bell Telephone Laboratories, Box 107, Red Bank, N. J.
Denison, J. W., 2026 Frederica St., Owensboro, Ky.
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Dietz, W. J., 4445 Vista St., Philadelphia 36, Pa.
Digan, T. E., Apt. 7G Springwood Village, Hyde Park, N. Y.
Dobren, S., 5902 Osage Ave., Philadelphia 43, Pa.
Drzewiecki, F. J., 235 Fifth St., Jersey City 2, N. J.
Edwards, A. H., 3953 Fairmount Ave., Philadelphia 4, Pa.
Eisenman, T. W., Box 0-133, Edwards Air Force Base, Calif.
Enoch, D. G., 351 Houston Ave., Mineola, L. I., N. Y.
Fabijan, F. M., Box 1121—3337 Technical Training Squadron, Scott AFB, Ill.
Feldstein, S., Box 315, APO 323, c/o P.M., San Francisco, Calif.
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Ferris, W. C., Jr., 2485 Woodscrest, Lincoln 2, Neb.
Flint, J. C., 802 South Painter Ave., Whittier, Calif.
Foxman, E., 4257 Kenyon Ave., Venice, Calif.
Gannon, G. F., Jr., 498 Campbell St., Pittsburgh 21, Pa.
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Gilat, J., 3 Zerubabel St., Haifa, Israel
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Gillespie, N. R., 8100 Monticello Ave., Skokie, Ill.
Gillette, A. J., Jr., 4005 Wesleyan Pl., San Diego 16, Calif.
Ginsburg, G. S., 2104 "O," Washington 7, D. C.
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(Continued on page 122A)

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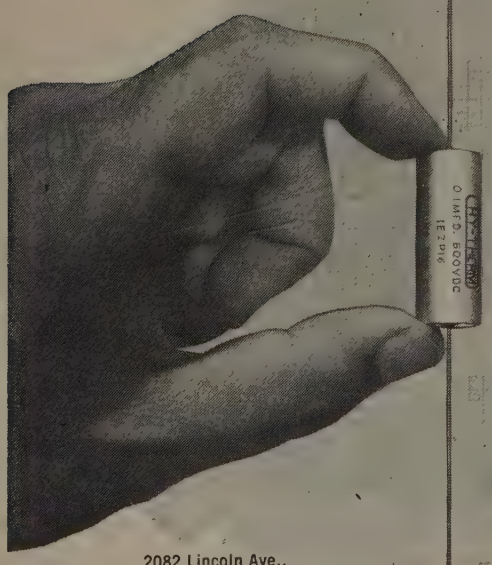

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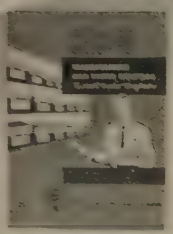

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Pennelli, A. P., 27 Brastow Ave., Somerville, Mass.
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Pilvick, C., 1254 Grant Ave., New York, N. Y.
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Pope, G. S., 1237 N. Brand Blvd., Glendale, Calif.
Quackenbush, H. H., 78-46 St., Sacramento, Calif.
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Rogers, D. W., 183-41 Babalon Ave., Hollis, L. I. N. Y.
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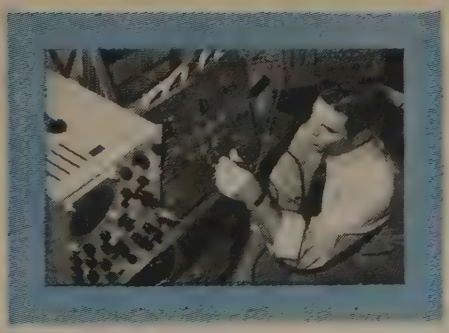

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(Continued from page 122A)

Sane, J. M., 8402 Kraay Ave., Munster, Ind.
Sawazaki, N., 3-1667 Minamissiwai-Cho, Kawasaki, Japan
Schmidt, A. F., Jr., RD 2, Washington, N. J.
Segalov, E., Box 7063 Hakirya, Tel-Aviv, Israel
Selditch, A. D., 345 Kansas St., Perkin Engineering Corp., El Segundo, Calif.
Shumaker, D. F., 601 E. Armour Blvd., Kansas City, Mo.
Simmons, G. J., 9201 Cherokee Rd., N.E., Albuquerque, N. Mex.
Spero, R. E., Sperry Gyroscope Co., Great Neck, L. I., N. Y.
Stambaugh, E. D., 3322 Soutter Ave., S.E., Cedar Rapids, Iowa
Stattleman, A. J., Electrical Engineering Dept., Iowa State College, Ames, Iowa
Stryker, E. M., Collins Radio Engineering Bldg., 5225 C Ave., N.E., Cedar Rapids, Iowa
Suzuki, Z., c/o Furukawa Electric Works, Nishi-ku, Yokohama, Japan
Swanekamp, J. E., 10312 Greenacres Dr., Silver Spring, Md.
Taylor, J. M., 56A Corlies Avenue, Poughkeepsie, N. Y.
Teal, H. E., 2457 Linwood Ave., Niagara Falls, N. Y.
Teegarden, W. M., 148 Grandview Ct., Ithaca, N. Y.
Theisner, W. G., 395 Page Mill Rd., Palo Alto, Calif.
Tomono, M., Hitachi Central Lab., Kokubunti-Machi, Kitatama-Gun, Tokyo, Japan
Tonik, A. B., 6245 Morton St., Philadelphia, Pa.
Tubergen, J. B., 2282 W. 11 St., Los Angeles, Calif.
Valleau, R. L., 1726 W. Ohio St., Chicago, Ill.
Van Collie, H. A., Powder Horn Hill, Wilton, Conn.
Velardo, J. M., 503 W. Loma Alta Dr., Altadena, Calif.
Wagner, R. G., 2450B 35 St., Los Alamos, N. Mex.
Walsh, J. L., Engineering Lab., Rm. N-122, Poughkeepsie, N. Y.
Watt, R. L., 2903 Lillian, Shreveport, La.
Weirich, H. P., 3158 S. 49 St., Milwaukee, Wis.
Wellsand, R., Jr., 1014 W. Oakdale Ave., Chicago 13, Ill.
Wikstrom, D. E., Rt. 1, Box 497, Brookmead Add., Bristol, Tenn.
Wolfe, G. M., 15514 Blackwood, Puente, Calif.
Wood, W. R., 242 Saratoga Dr., San Antonio, Tex.
Wuster, W. L., 1930 N. Leland St., Indianapolis, Ind
Yoakum, W. E., Rt. 2, Box 676, Huntsville, Ala.
Zilmer, D. E., 109A Richmond Rd., USNOTS, China Lake, Calif.

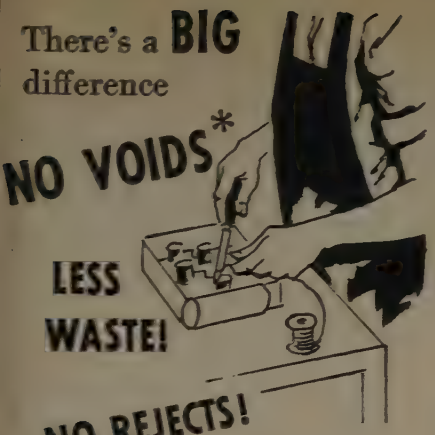
SECTION MEETINGS

ATLANTA

Tapescript: "Physics of Music and Hearing" by Dr. W. E. Cock, Bell Telephone Labs.; June 18, 1954.

(Continued on page 126A)

There's a **BIG** difference



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INPUT IMPEDANCE: 1 Megohm shunted approx. 20 mmF for each single-ended or differential input; 2 Megohms shunted approx. 10 mmF for push-pull signals

FREQUENCY RESPONSE: d.c. to 100,000 cycles within 1 db, 6 db down at 200,000 cycles

MAXIMUM GAIN: 100 ± 10% for balanced or unbalanced inputs or outputs

INPUT ATTENUATORS: 100, 10 or 1 and "off" positions, independently in each channel

DC INPUT CONNECTIONS: Two posts for push-pull or differential, one post connected to ground: for single-ended signals, other input is disabled by setting attenuator to "off" position

AC INPUT CONNECTIONS: Arranged like DC connections; signal terminals connected to DC terminals through coupling condensers

OUTPUT IMPEDANCE: Approx. 250 Ohms single-ended, approx. 120 Ohms push-pull. Minimum load resistance; 250 Ohms

UNDISTORTED OUTPUT: 20 volts or 5 Milliamperes (peak) max.

NOISE AND HUM: Below 40 microvolts referred to input

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POWER CONSUMPTION: 120 watts from 115 volt 50-60 cycle, single-phase a.c. power line

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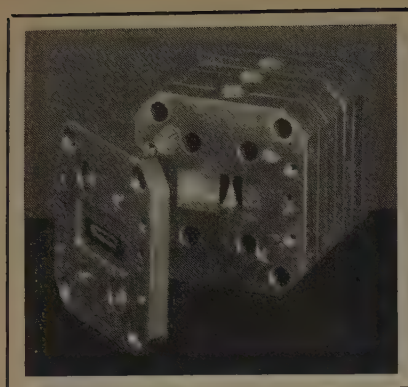

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SECTION MEETINGS

(Continued from page 124A)

BALTIMORE

"Characteristics and Applications of Microstrip for Microwave Wiring," by Maurice Arditi, Federal Telecommunication Lab.; September 8, 1954.

COLUMBUS

"Pulse Code Modulation and Its Application to Telemetering Systems," by B. D. Smith, Jr., Melpar, Inc.; April 6, 1954.

Inspection Tour of North American Aviation's Columbus Plant; April 28, 1954.

"The Functional Organization of Digital Computers," by James Stone and Bruse Moncreiff, Battelle Memorial Institute; May 18, 1954.

"Seward's Folley" by R. J. Anderson, Battelle Memorial Institute; June 9, 1954.

CONNECTICUT VALLEY

"Vibration Analysis," by R. C. Lewis, Calidyne Company; September 16, 1954.

DAYTON

"Reliability Consideration in the Design of Guided Missle Test Equipment," by J. F. Tobias, Hycon Mfg. Company; September 2, 1954.

EL PASO

"Remote Control and Radio Controlled Servomechanisms," by J. H. Straley, Radiophone Company; August 27, 1954.

EVANSVILLE-OWENSBORO

"The Role of Flight Directors in Present Day Aircraft," by N. L. Graham, Collins Radio Company; and hobby talk: "Color Photography," by J. R. Somerville; September 8, 1954.

HUNTSVILLE

"Automation," (Miniaturization of Electronic Assemblies, and Automatic Manufacture) by S. G. Bassler, Signal Corps Engineering Lab., Fort Monmouth; August 24, 1954.

INYOKERN

"Dielectric Absorption in Capacitors," by E. Malloy, Beckman Instrument Company; September 13, 1954.

LITTLE ROCK

"Nationwide Long Distance Dialing," by J. H. Landwehr, Southwestern Bell Telephone Company; September 14, 1954.

LOS ANGELES

"What Is the West Coast Manufacturer's Association," by E. P. Gertsch, Gertsch Products; "Theory of Failure," by R. Weller, Naval Air Missile Test Center, Pt. Mugu; and "The Use of Electronics in Civil Defense and Municipal Governments," by M. E. Kennedy, Director of Communications, Los Angeles County; September 7, 1954.

(Continued on page 128A)

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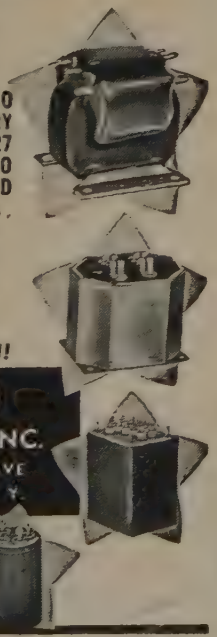

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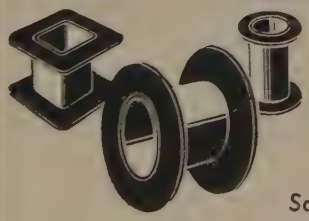

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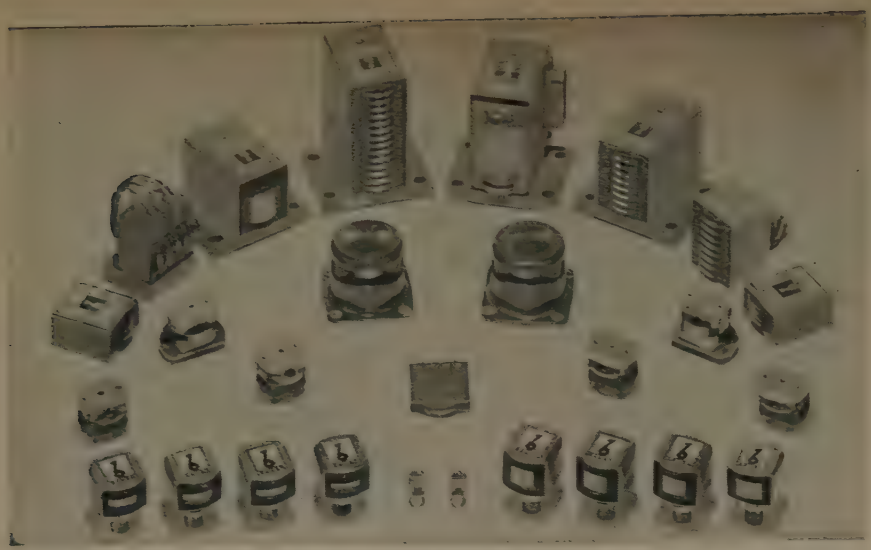

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SECTION MEETINGS

(Continued from page 126A)

NORTHERN NEW JERSEY

"Silicon in Modern Communications," by G. L. Pearson, Bell Tel. Labs.; and "Grown-Junction Silicon Transistors," (by M. Tannenbaum and L. B. Valdes) presented by Mr. Valdes, Bell Tel. Labs.; September 15, 1954.

PITTSBURGH

"Radio Studies of the Sun," by W. L. Roberts Westinghouse Research Labs., September 13, 1954.

PORTLAND

"Development of High Voltage Supplies and Regulators," by Dr. Victor Wouk, Beta Electric Corp. August 18, 1954.

ROME-UTICA

"Ideas of Modern Physics," by Dr. Paul Weiss, General Electronics Lab.; September 7, 1954.

SACRAMENTO

"The Vidicon and Its Applications," by John Day, Kay-Labs.; September 10, 1954.

SAN DIEGO

"Recent Advances in Microwave Tubes," by Dr. J. R. Pierce, Bell Telephone Labs.; August 30, 1954.

SAN FRANCISCO

"Advances in Microwave Tubes," by J. R. Pierce, Editor, PROCEEDINGS OF I.R.E. September 7, 1954.

SYRACUSE

"What's With UHF Television?"—panel discussion by P. L. Chamberlain, Dr. L. R. Fink and C. J. Simon, all of General Electric Company; September 9, 1954.

TOLEDO

Film "Atomic Energy"; September 9, 1954.

WASHINGTON

"Color TV-Derivation of NTSC Standards," by J. W. Wentworth, TV Terminal Equipment Engineering; September 13, 1954.

SUBSECTIONS

MONMOUTH

"Speech Transmission in Binary Form," by A. H. Ross, Signal Corps Eng'g. Labs., Fort Monmouth; September 15, 1954.

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New Panoramic engineering achievements embodied in these improved instruments open important new applications involving modulation and interference problems.
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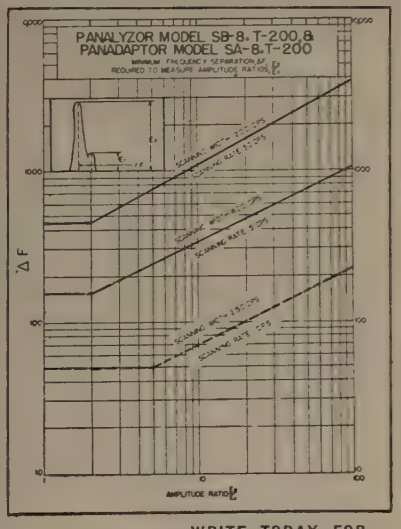
3 types available with maximum sweepwidths of 200 KC., 1 MC. and 10 MC.



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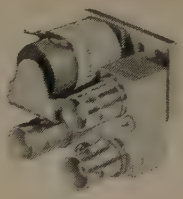
FOR SWITCHING COAXIAL CIRCUITS AT FREQUENCIES UP TO 5000 MC, DANBURY-KNUDSEN OFFERS A TYPE FOR ALMOST ANY PRODUCT APPLICATION

SPDT Coaxial Switches with grounding-type contacts are available in series 4 and 5 with many combinations of MB, SM, BNC, UHF, N, and C connectors, as well as attached lengths of cable, or solder terminals. Both series come in either 100-watt size or a larger size which has a power rating of 500 watts and twin contacts in both positions. **The VSWR** of these switches is low. Average measurements of the CR25N (N connectors) show VSWR readings ranging from 1.03 at 100 mc to 1.4 at 4000 mc. Over the same frequency range, the crosstalk figure varies from 85 db to 40 db.

For Limited Space Applications, miniature SPDT coaxial relays are available in series 2. These relays have MB or BNC connectors or attached lengths of cable. $1\frac{5}{16}$ inches high, relays in this series require a mounting space of only $1\frac{5}{16}$ by $1\frac{3}{8}$ inches.



Double-Pole Transfer (crossover) switches are available in series 56 with either BNC or N connectors. At 300 mc, VSWR is 1.08 and the crosstalk figure is 44 db.

Fast Operating Multiposition switches are available in 3-, 4-, or 6-position types with BNC connectors, and in 3- and 4-position types with N connectors. At 500 mc, VSWR is 1.1 and crosstalk 60 db.

Resistor-Terminated Switches for video and similar applications are available in 2-, 3-, and 4-position types with BNC connectors, and in 2-position types with UHF or N connectors. Each input is terminated in a 50 or 75 ohm resistor, when not connected to the common output.

Small Rotary Switches in 3-, 4-, or 6-position types with BNC connectors constitute series 1000. These switches are 2 inches in diameter, manually operated, with a VSWR of 1.15 and crosstalk figure 60 db (both measured at 1000 mc).

Most of the 2-position switches above are available for either AC or DC remote operation, or for manual operation by means of a toggle type actuator. For remote operation, the multiposition types can be supplied only with DC operating coils.

Mating connectors for all cable types may be obtained from the INDUSTRIAL PRODUCTS COMPANY, a division of DANBURY-KNUDSEN, INC.

Write for our catalog describing these and other types of coaxial switches and relays.

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The New JONES FANNING STRIP

The correct wire to correct terminal every time!

Connections are made through Fanning Strip, on bench or anywhere apart from barrier strip, and quickly slipped into assembly.

Designed for use with Jones Barrier Terminal Strips Nos. 131 and 142, for 1 to 20 terminals.

Simplifies and facilitates soldering. Insures positive correct connections. Saves time. Ideal for harness or cable assembly. Strong construction: Brass terminals, cadmium plated. Heavy bakelite mounting.

9-141 Barrier Strip

1-181 Fanning Strip. Pat. applied for.

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- OVER 15 MC. BANDWIDTH
- MILLIMICRO-SECOND ACCURACY
- EXCELLENT TRANSIENT RESPONSE

TYPE 505

MINIATURE CONTINUOUSLY VARIABLE DELAY LINES

Type	Delay	Impedance	Type	Delay	Impedance
505	0 to 0.05 us	330 ohms	506c	0 to 0.32 us	220 ohms
505a	0 to 0.1 us	180 ohms	506d	0 to 0.37 us	240 ohms
506	0 to 0.25 us	200 ohms	507	0 to 0.75 us	580 ohms
506a	0 to 0.17 us	180 ohms	507a	0 to 0.5 us	185 ohms
506b	0 to 0.28 us	190 ohms	508	0 to 0.4 us	580 ohms

$59.00 EACH — SIZE 4" x 4" x 1"

CONTINUOUSLY VARIABLE PASSIVE NETWORK. Capable of providing continuously variable time delay from 0 to 11 microseconds with 5 x 10^{-10} Sec. resolution time. Type 2011—$245 each. Type 303 (11 models) —$144 each.

STEP VARIABLE DELAY LINES. Equal input and output impedance available with maximum delay of 2.2 us to 10 us.

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News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 33A)

Encapsulated Precision Resistors

A new line of Riteohm ® encapsulated, wire-wound precision resistors has just been announced by **Ohmite Manufacturing Co.**, 3601 Howard St., Skokie, Ill. These resistors exceed the requirements of specification MIL-R-93A, Amendment 2, at 125°C ambient.

The enameled alloy resistance wire is noninductively pie-wound on a steatite bobbin, which holds the pie windings in place and resists mechanical movement. The resistance wire is welded to the terminals. The entire assembly is encapsulated (including center of bobbin) in a special resin. This resin has a coefficient of expansion that closely matches that of the bobbin, wire, and terminal materials.

The new Riteohm ® resistors are available with axial leads (Type 85) or lug terminals (Type 86). Wattage sizes are $\frac{1}{4}$, $\frac{1}{3}$, $\frac{1}{2}$, $\frac{3}{4}$ and 1 watt (125°C ambient). Resistance values are available up to a maximum of 1,950,000 ohms. Resistance tolerances as close as ± 0.1 per cent can be furnished.

Axial-lead type can be furnished in eight sizes from $\frac{3}{8}$ to 1 inch in length (excluding leads) and from $\frac{5}{16}$ to $\frac{1}{2}$ inch in diameter; the lug type can be furnished in three lengths from $\frac{1}{2}$ to 1 inch and $\frac{5}{16}$ inch in diameter.

Bulletin available.

Instrument Generator

A nine-ounce, permanent-magnet, continuous-duty generator recommended as a precision voltage source is described in a new leaflet printed by **Dalmotor Co.**, 1360 Clay St., Santa Clara, Calif. Part of a series, the leaflet, Form GPM-44A, illustrates the unit, provides dimensional outlines, and tabulates physical and electrical characteristics. Operating parameters of phase voltage vs. load-per-phase are presented graphically for operating speeds of 1200, 2100, and 4500 rpm, giving frequencies of 20, 35, and 75 cps.

(Continued on page 132A)

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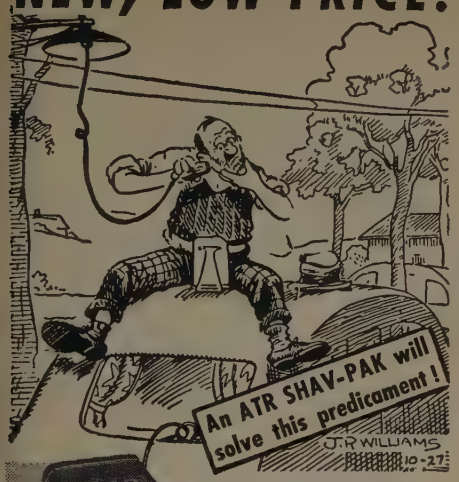

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TYPE	INPUT D.C. VOLTS	A.C. OUTPUT 60 CYCLES	OUTPUT WATTAGE	LIST PRICE
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12-SPB	12	115	15	995

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WITH GREATER OUTPUT

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- WIRE RECORDERS

TYPE	INPUT D.C. VOLTS	A.C. OUTPUT 60 CYCLES	OUTPUT WATTAGE	LIST PRICE
6-DME	6	115 volts	30-40	19.95
6H-DME	6	115 volts	60-75	29.95

Above Inverters also available for 12-volt operation.

See your jobber or write factory today for complete information

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Quality Products Since 1931
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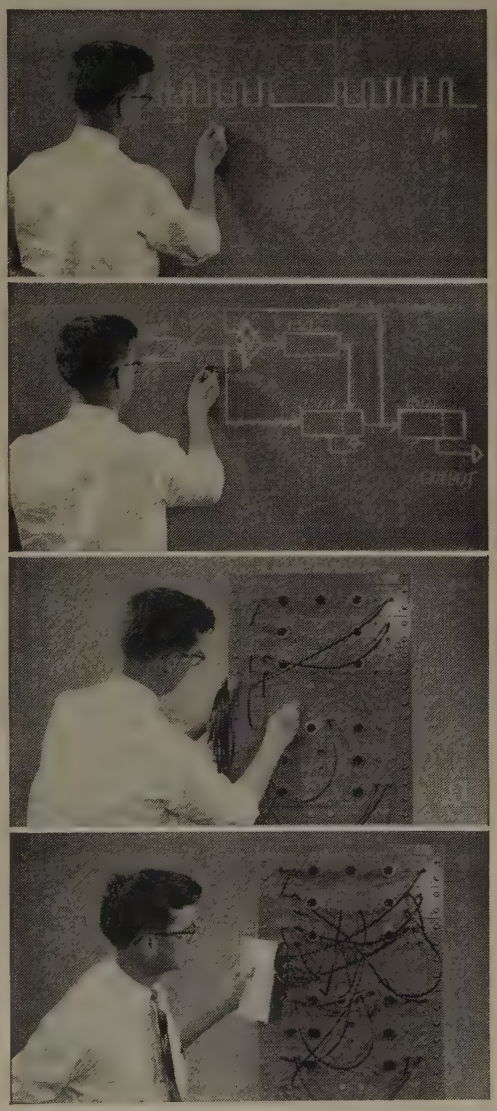
Engineer assembles pulse system in 30 minutes, using Burroughs "do-it-yourself" units

Standard, matched units, performing basic functions, connect together to form even the most complex pulse systems



1. START
Engineer studies time chart of the desired pulse sequence. This is the output he wants the pulse system to produce.

2. PLANNING THE SYSTEM
He determines which Burroughs units he needs and how they should be connected together. This can be done by means of a simplified block diagram. Time: 10 minutes.

3. ASSEMBLING THE UNITS
Using standard coaxial cables, he completes his pulse system by connecting the units together according to his block diagram. Time: 20 minutes.

4. JOB COMPLETED
System now produces the exact pulse sequence desired. Engineer saves weeks of breadboard engineering, vital time, uncertainty, and considerable equipment cost. And his Burroughs "Unitized" pulse handling equipment can be used over and over again on different future projects.

GET THE FACTS

No matter how complex the pulse sequence you need, you can produce it quickly and at relatively low cost with Burroughs "Unitized" pulse handling equipment. If you prefer, send us a timing diagram of the pulse sequence required, and we'll advise you what Burroughs units you need and the cost. Immediate delivery from stock. Write Burroughs Corporation, Electronic Instruments Division, Dept. 11 L, 1209 Vine St., Phila. 7, Pa.

ELECTRONIC INSTRUMENTS DIVISION
Burroughs
FIRST IN PULSE HANDLING EQUIPMENT

HIGHEST STANDARDS
TOP QUALITY!
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"CONDENSER SPECIALISTS SINCE 1923"
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Precision gauges from ⅛ to .002.
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• Accelerating potential of 4 KV
• Illuminated graticule
Write for complete technical specifications
LABSCOPE inc.
728 GARDEN STREET · CARLSTADT, N.J.

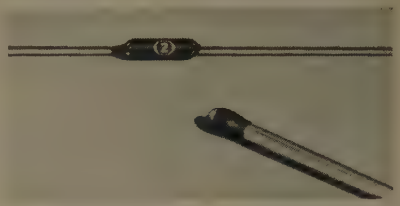
News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 130A)

Miniature Wire-Wound Resistors

A new subminiature 3 watt wire-wound resistor which is the same size as conventional ½ watt molded carbon resistors is now available from the **Sprague Electric Co.,** 235 Marshall St., North Adams, Mass.



Developed especially for use in military and industrial electronic equipment, this unit, approximately the size of a match head, is expected to find application in point-to-point and terminal board wiring as well as on printed wiring boards. The new unit is 13/64 inches in length and has a maximum resistance value of 10,000 ohms. The resistor is provided with blue vitreous enamel coating which will withstand stringent humidity tests.

Complete description of this new resistor is given in Engineering Bulletin 111-B, which lists standard resistance values.

Speed-Regulated Motor

Providing a starting torque of 110 pound/inches in a continuous-duty unit, the new Type SC-5 Motor manufactured by **Dalmotor Co.,** 1360 Clay St., Santa Clara, Calif., produces 1 hp at 12,000 rpm, regulated to ±6 per cent in speed variation. It draws 45 amperes, operating over the dc voltage range from 25 to 29.5 and regulates over a load variation of ⅔ to 1 hp.

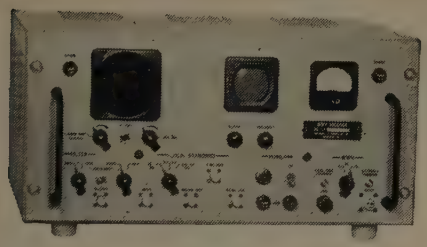
Bypass capacitors are built in at the armature and governor brushes to assist in the attainment of minimum radio interference in operation. Input connections are by means of a shielded positive lead and a grounded negative stud, although special arrangements can be supplied where required.

The unit has a geared output shaft, but can be supplied with a plain shaft of length as required, or splines, keyways, or other such arrangements. Weighing 10.8 pounds, the SC-5 is standard with counter-clockwise rotation when the shaft end is viewed.

(Continued on page 134A)

WWV IS FREE

Why Not Use It?



The WWV frequency standard, as transmitted by the U.S. Bureau of Standards, is the most accurate frequency standard of any in the world.

By using the new Matawan WWV receiver, this frequency standard, with all its information (standard audio frequencies, time signals, propagation notices) is now available to your laboratory. The WWV carrier, after elimination of its modulation, by means of a narrow band pass crystal filter, monitors your local HF standard. The rotative pattern on the screen of the cathode tube gives sense and direction of the frequency deviation.

Either the 5MC WWV or the 15 MC WWV may be received. The receiver will accommodate any local frequency in the range from 50kC to 10MC equal to a subharmonic of 10MC, 2MC or 400 KC. A second channel receives the audio modulation and time signals of the WWV-Standard.

Four models available.

	WWV Input 5 MC	5 and 15.0 MC
Audio Modulation on Speaker Phone Jacks	Model ME-117A $865.00	Model ME-117C $910.00
Additional Filter for 440/600 CPS	Model ME-117B $940.00	Model ME-117D $985.00

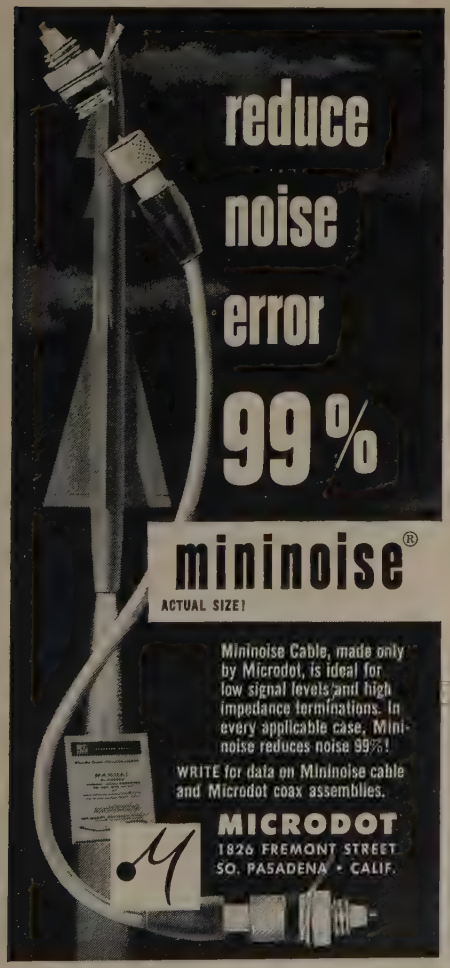
F.O.B. Matawan, N.J.

Phillip H. Harrison & Co.

Route 34, Matawan, N.J.

Matawan 1-3113, 1-4242

Write for Illustrated Descriptive Pamphlet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B&W WINDS THEM *ALL!*

Now . . . you can have a dependable source of supply . . . for toroids of *any* size . . . in small or large production runs.

B&W's recent purchase of one of the first *Boesch Subminiature Toroidal Coil Winding Machines* extends to an even greater range the complete line of B&W toroidal coils available to meet any of your requirements. For with this new equipment, B&W can now wind the smallest toroidal coils on a production line basis, just as precisely, just as rapidly, as it always has the larger sizes.

The ability of these miniature coils to reduce the size of many different types of toroidal assemblies is nowhere better illustrated than in the company's own line of filters. Made to meet the highest standards of industry and military, these B&W components include high-pass, low-pass, band-pass, and band suppression filters, as well as discriminators for r-f or audio frequencies.

For information on how B&W toroids can solve your production problems, give you greater miniaturization, write us about your problem on your company letterhead.

B&W **Barker & Williamson, Inc.**

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Serving Industry—FOR OVER 53 YEARS
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Very small diameter—for filaments, thermo couples, resistance units. Available bare or enameled.
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Produced in Platinum, Gold, alloys and pure metals—small diameter. Platinum alloy resistance wire bare or enameled.
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Comprising an extensive range of wires, electroplated with various metals, such as Gold, Silver, Rhodium, Platinum, Copper, Gallium, Indium and others . . . Enamel insulated wires for precision resistors and potentiometers.
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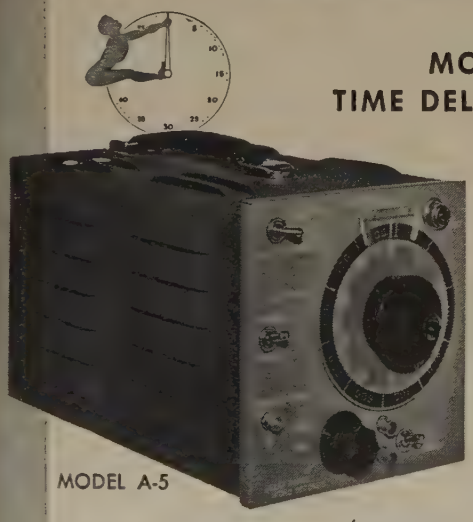


MODEL A-5

MODEL (A-5)
TIME DELAY GENERATOR

An instrument of small size which provides accurate and variable time delays from 1 μs to 1,000 μs in three ranges.

• low jitter (.008%) • linear scales • highly stabilized power supplies • small repetition rate effects • blocking oscillator output • trigger type input • separate calibration adjustments for each range.

Get complete data: our Bulletin I-A-5

Other Rutherford precision Time Delay Generators { Model A-2 (.8 μs to 100,000 μs): Bulletin I-A-2 / Model A-4 (10 μs to 10 secs): Bulletin I-A-4

Rutherford ELECTRONICS CO.
Telephone: TExas 0-4362
3707 S. ROBERTSON BLVD.
CULVER CITY, CALIFORNIA

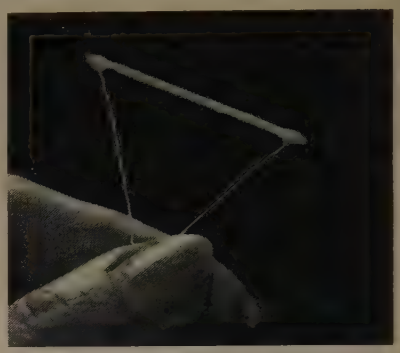
News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 132A)

Thermistors for Radiosons

A new thermistor designed specifically for temperature measurement and control in the range of −60° C. to +40° C. is now available from the **Ferroxcube Corp. of America,** 235 E. Bridge St., Saugerties,



N. Y. This temperature sensitive resistance element, in the form of a thin toothpick-like rod with radial leads, is made of metal oxides and manufactured by a ceramic process. All units are waterproofed and coated with reflective coating. Performance meets the requirements of military specification MIL-T12284a (Sig. C.).

(Continued on page 136A)

Measurements Corporation
MODEL 79-B

PULSE GENERATOR

REPETITION RATE: 60 to 100,000 pulses per second.

PULSE WIDTH: Continuously variable from 0.5 to 40 microseconds.

RISE TIME: Approx. 0.25 microseconds.

OUTPUT VOLTAGE: 150 volts (peak) positive.

POWER SUPPLY: 117 volts, 50/60 cycles.

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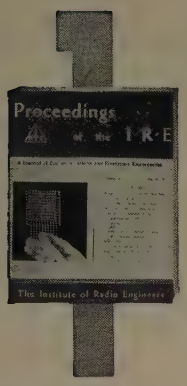

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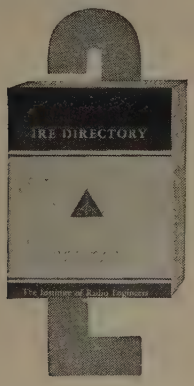
puts your product promotion monthly before the "thinking and doing" engineers in the fabulous, fast-moving radio-electronic industry. Circulation 41,625 (ABC).

Founded in 1913, "Proceedings of the I.R.E" is the authoritative work-book of America's fastest expanding industry. In it, electronic advances, years ahead of the times, are published every month, thus keeping engineers up to the moment on all that is new in engineering research and application. Its advertising is a constant product parade of the best this industry manufactures. A vital service to all men who specify and buy, it is a proved vehicle for your consistent product promotion—*with prestige!*



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Bringing buyers and sellers face to face to discuss engineering needs and product performance is one of IRE's great services to an industry where understanding is the key to progress. The Radio Engineering Show completes with actual product presentation the balanced promotion package of "Proceedings of the I.R.E" for product promotion and the Radio Engineers Directory for product reference.

For complete facts, ask IRE about all 3!

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PULSE DISTRIBUTION AMPLIFIER

Wickes

MODEL PDA-2

Non-inverting pulse distribution and isolation amplifier. High input impedance for bridging directly across signal lines. Two input receptacles for looping signal lines, mixing, or terminating. Six identical, isolated output channels, each designed for operation into a 75-ohm load. Master gain control, with overall gain adjustable from 1.0 to 2.0 Frequency response ±1 db from 50 cps to 10 mc. Adjustable high-frequency peaking. Passes 50 cps square wave with less than 5% tilt. Built-in, fused filament supply and regulated bias supply. Supplied with six terminations. For standard rack mounting.

Write for data sheets on our complete line of color television instruments for signal generating, distribution, test and measurement, and signal certification.

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SPECIMEN PANEL MIL-P-7788 (AN-P-89) SENT ON LETTERHEAD REQUEST

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 134A)

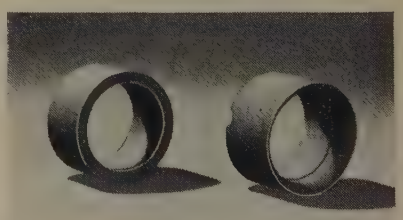
Cardiac Stimulator

Producing closely regulated but adjustable serial pulses, the Morris Clinical Pacemaker available from **Levinthal Electronic Products, Inc.,** 2889 Fair Oaks Ave., Redwood City, Calif., provides an output suitable for the control of heart beat in cases of cardiac standstill or conduction block. The unit can be used externally with either electrocardiograph-type chest-strap electrodes or with hypodermic needles as electrodes; or, in open-chest cases, in direct contact with the ventricles.

Ready for full operation within five seconds after being energized, the instrument can be used simultaneously with an electrocardiograph without interference. There is a minimum number of simple controls, and the circuitry includes isolation transformers and other special safety features.

Pulse output is controllable for a wide range of applications. The monophasic positive rectangular round-top pulses of two milliseconds duration are continuously variable from 0- to 130-volt amplitude. Heart rate, which is fully automatic, is continuously variable from 50 to 150 pulses per minute. Requirements 115 volts, 60 cps. Price is $375.

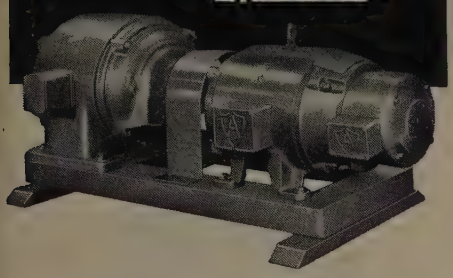
Ferrite Yoke Cores



Two new microgap ferrite deflection yoke cores for use in television receivers are now available from the **Ferroxcube Corp. of America,** 235 E. Bridge St., Saugerties, N. Y. Type 200T1250B201 is a 70° core with improved performance characteristics, while Type 189T1438B201 is the corresponding 90° core which has been developed by Ferroxcube. Both cores are made with Ferroxcube 3C manganese zinc ferrite material. The 90° core is at the right in the illustration.

(Continued on page 138A)

Are your High-Cycle Alternator requirements *special?*



Where lowest possible maintenance, combined with excellent wave form characteristics are required, it's hard to match American Electric's Inductor-Type Alternators. But occasionally *special requirements arise* which may be better served by American Electric's Revolving Field Alternators.

Here Are Quick Facts:

AMERICAN ELECTRIC REVOLVING FIELD ALTERNATORS

CAPACITY RANGE—15 KVA to 40 KVA (in stock). Up to 75 KVA on special order.

FAST RECOVERY—Better than .2 seconds.

LOW VOLTAGE OVERSHOOT—Less than 10%.

TOTAL HARMONIC CONTENT—Under 5% on unbalanced 3-phase loads

These characteristics apply to general uses. Where applications are specific, even better characteristics can be developed.

EXCITER REGULATING SYSTEM

Employs American Electric's trouble-free, direct-connected, high-cycle Inductor Alternator—which has no commutator, slip rings, brushes, springs etc. Output is rectified and exciter-regulated by either electronic or magnetic amplifier means then returned to the rotating field of the main alternator.

Write for details and quotations outlining requirements.

Also Manufacturers of High Frequency Inductor Alternators, Miniature Electric Motors, A.C. Industrial Motors, Motor Driven Blowers and Fans.

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HIGHER CAPACITY OR • HIGHER FREQUENCY •

WHATEVER THE DESIGN REQUIREMENT

for vacuum fixed or variable capacitors, JENNINGS can meet the need with a large selection of capacities, sizes, and voltage ratings.

High capacity units up to 5000 mmfd at 10 kv and 1000 mmfd at 45 kv are now being made in both variable and fixed types.

Small vacuum capacitors such as the one shown above are currently being used in VHF and UHF applications. The ECS, for example, has capacity ranges of either 2 to 8 mmfd or 3 to 30 mmfd and a voltage rating up to 15 kv.

Send for our catalog summary describing over 300 types, sizes, and ratings of vacuum capacitors and switches.

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where *Quality* counts
"F" Series RACKS by PAR-METAL
No. F-6618: Overall: 67⅜" x 22" x 18"
Panel Space: 61¼" x 19"
No. F-8318: Overall: 83⅛" x 22" x 18"
Panel Space: 77" x 19"
Want "custom-quality" in a standardized, economical Rack? Here's the industry's top value.
- Front and rear doors.
- Streamlined, modern in design. Ruggedly constructed of furniture steel.
- Entire cabinet is welded together into an integral unit.
- Note: full adjustment of panel mounting angles is provided for, to position material at any distance from front or rear doors.
- Finishes: black ripple, grey ripple, aluminum grey lacquer (or prime coat only).

Planning an electronic product? Consult Par-Metal for
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Remember, Par-Metal equipment is made by electronic specialists, not just a sheet metal shop.
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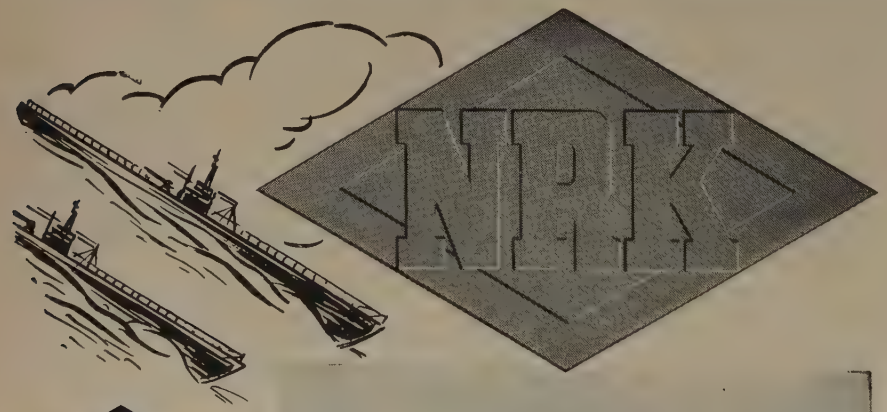

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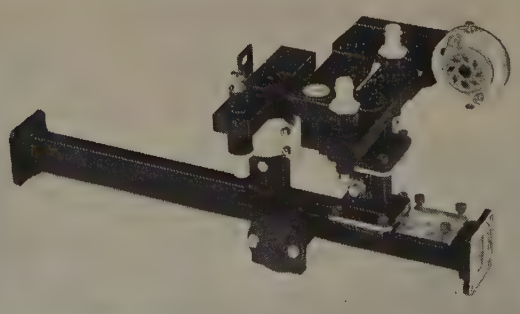

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News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 136A)

Envelope Delay Tracer

Wickes Engineering and Construction Co., 12 St. & Ferry Ave., Camden 4, N. J., has a new envelope-delay tracer. It is a precision instrument designed to measure the envelope-delay characteristic of amplifiers, phase correction, and distribution networks, complete color or monochrome television systems. The unit will also determine the amplitude characteristic of the equipment under test.

The EDT-1 does away with plotting and mathematical computation of delay. In use there are no "dead" spots in the frequency range. Delay can be measured at any desired frequency over the video range of 0.5 mc to 6.0 mc The envelope delay meter is calibrated to read delay from zero to ± 1.25 μs. The scale is calibrated in 0.05 μs increments from 0.3 to 1.25 μs. A separate meter is provided for determination of the amplitude characteristic.

Measurements can be made point-by-point using single frequencies, or the envelope delay can be measured over a desired band of frequencies using a sweep generator and oscilloscope.

Telelink Bulletin

Announcement has been made by **Raytheon Manufacturing Co.,** Waltham 54, Mass., of the availability of an 8-page 2-color brochure describing the company's Model TCR-12 "Telelink" for use in microwave communications systems.

While the Raytheon "Telelink" will be used for the extension of rural telephone service, the new equipment will also be used for microwave communications systems, pipelines, railroads, utilities, air lines, turnpike authorities, and federal, city and state governments.

Full specifications of the equipment are included.

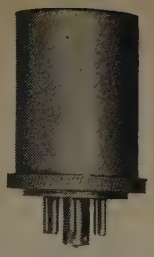
Copies of the brochure may be obtained by writing for literature No. DL-TL-912, Raytheon Manufacturing Company, Dept. 6130, 100 River St., Waltham 54, Mass.

(Continued on page 140A)

VHS* RELAY

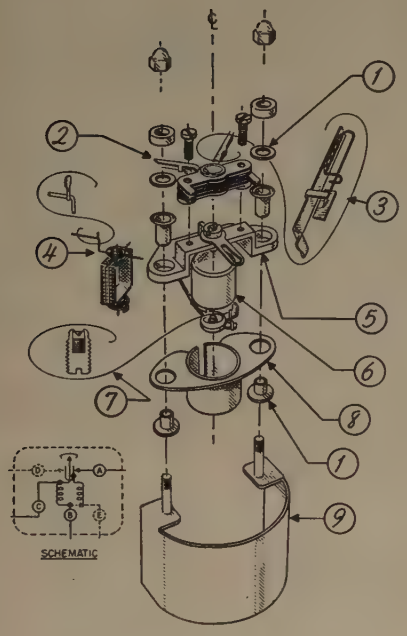
(*Very High Sensitivity)

The VHS is a balanced armature, Alnico magnet type relay. It is internally shock mounted and resistant to vibration. The screw-on cover is gasket sealed. It can be opened and resealed.



0.2 microamperes, (3000 ohms coil) or, 0.1 millivolts, (20 ohms).

Connections: 9 pin octal style.
Dimensions: 1¾" dia. x 2¼" long.
Weight: 4 ounces.



1. Shock mount
2. Contact assembly
3. Contact detail
4. Armature with contact detail
5. Cast bracket
6. Alnico magnet
7. Bearing detail
8. Yoke (steel)
9. Mounting frame

Sensitivity: Infinite variations from 0.2 Ua. to 10 Amp. or 0.1 Mv. to 500 volts, self contained. Higher volts or amps with external multipliers. A.C. rectifier types. Trip point accuracies to 1%. Differential and repeatability better than 1%.

Contacts: SPST or SPDT, 5-25 Ma. D.C. Other ratings to ½ Amp. A locking coil gives high pressure and chatter free contact even under shock and vibration. Samples 3 to 4 weeks.

ASSEMBLY PRODUCTS, Inc.

CHESTERLAND 2, OHIO

Ask for Bulletin 102

Introducing— THE MYCALEX PARTS KIT

✳ For design projects

✳ For experimental use

✳ For emergency repairs

To acquaint engineers with the superior qualities of Mycalex glass-bonded mica—the world's most nearly perfect insulation—a conveniently packaged kit containing more than 30 Mycalex products has been assembled.

This assortment includes 46 miniature and sub-miniature tube sockets, terminal boards, rods and strips of fabricated Mycalex 400 and various Mycalex 410 injection-molded products such as connector sleeves, coil forms, stand-off terminals, rotors and rings, motor slot wedges and switch wafers. A handy reference table compares the superior physical and electrical properties of Mycalex glass-bonded mica with those of other insulating materials.

MYCALEX

✳ withstands extreme operating temperatures

✳ offers high arc resistances

✳ possesses low loss and power factors

✳ for permanent dimensional stability

The Introductory Mycalex Kit is being offered to qualified engineers and others who are interested.

Mycalex Corporation of America, Dept. III
Clifton Boulevard, Clifton, New Jersey

Please send me the new Mycalex Parts Kit at the introductory price of $10.00.

Enclosed find Check.....Cash.....Money Order.....

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Company.....

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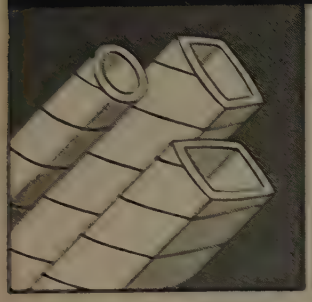
MYCALEX CORPORATION OF AMERICA

World's largest manufacturer of glass-bonded mica products

EXACTLY your type and size
in Spiral Wound PAPER TUBES



SQUARE, RECTANGULAR, ROUND

Standard-type PARAMOUNT paper tubes used for millions of coil forms and other applications. Hi-Dielectric. Hi-Strength. Kraft, Fish Paper, Red Rope, Acetate, or any combination wound on automatic machines. Any size from ½" to 30" long, from .450" to 25" I.P. Produced from wide range of stock arbors or specially engineered for you.

NEW "PARAFORMED" TUBES
SQUARE OR RECTANGULAR

Entirely new technique in tube making developed and perfected by PARAMOUNT. Perfectly flat side walls, sharp square inside corners, and very small radius on the four outside corners. Spiral wound, not die formed. No sharp outside edges to cut wire. No need for wedges to tighten winding on laminated core. Full rigidity and physical strength. Permits winding coils to closer tolerances. Allows faster automatic stacking of coils. Approved and used by leading manufacturers. *No extra cost!*

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Complete with all five Tuning Units, covering the range 38 to 4,000 Mc.; wideband discone and other antennas, wavetraps, mobile accessories, 100 page technical manual, etc. Versatile, accurate, compact—the aristocrat of lab receivers in this range. Write for data sheet and quotations.

We have a large variety of other hard-to-get equipment, including microwave, aircraft, communications, radar; and laboratory electronics of all kinds. Quality standards maintained. Get our quotations!

NEW TS-13/AP X-BAND SIGNAL GENERATORS, with manual, $850.00 . . . TS-175/U Frequency Meters, 8S-1,000 Mc., $625.00 . . . T-47/A ART-13 Transmitters, $450.00 . . . and many more!

ENGINEERING ASSOCIATES
434 PATTERSON ROAD
DAYTON 9, OHIO

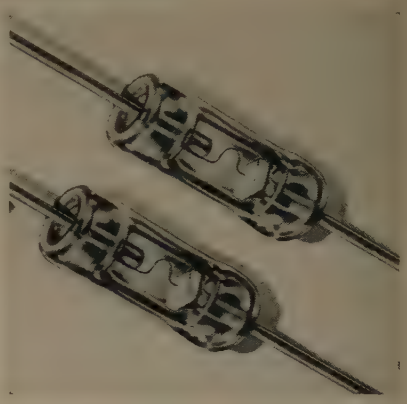
News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 138A)

Subminiature Glass Germanium Diodes

Transitron Electronic Corp., Dept. P., 403 Main St., Melrose 76, Mass., has developed several new subminiature glass germanium diodes.



There are computer, high temperature, high resistance, high conductance, and general purpose types available. The firm has complete data sheets prepared on each of these types.

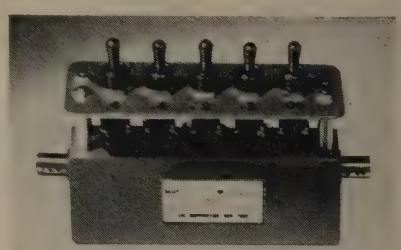
All are hermetically sealed in a one piece glass case. The tinned dumet lead is permanently fused to the germanium wafer, and the gold wisker wire is bonded to the germanium. The dimensions are 0.125 inch maximum diameter, 0.265 inch maximum length, and 1.25 inches minimum lead length.

The components are tested for shock, vibration, and temperature and humidity cycling between −78° and +135° C.

One type, T8G offers over 5 megohms at 100 volts inverse, and under 10 ohms at 1 volt forward. Request bulletin TE-1319.

Delay Line

ESC Corp., 534 Bergen Blvd., Palisades Park, N. J., has developed a new panel mounting delay line which has an adjustable delay in steps of 0.05 μsf to a maximum of 1.15 μsf and a characteristic impedance of 72 ohms.



Pulse rise time is 0.13 μsf after a total delay of 1.15 μsf and nominal delay tolerance is 3 per cent.

Input and output connections are through BNC jacks. The unit is hermetically sealed with the network embedded in epoxy resin. Overall dimensions excluding jacks and switches are $4\frac{5}{8} \times 1\frac{3}{4} \times 3\frac{1}{4}$ inches deep.

(Continued on page 142A)

color TV shadow masks

We are now delivering shadow masks, the precision heart of the color television tube, etched to specifications. Mass production methods, perfected during more than 40 years of precise metal etching experience, enable us to meet today's exacting demands for quality and delivery.

etched and electro formed precision parts

With the technical experience and production know-how gained in producing fine metal etching and in electro-forming, we are now manufacturing many components for the electronic industry. Some specifications call for fine mesh up to 1,000,000 holes per square inch. Send us your drawings for a prompt quotation.

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- Provides accurate test data
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Latest CUBIC designs provide vector position coordinate data in non-ambiguous digital form. Simplified portions of the system permit recording of slant range or miss distance on a special phase meter. Multiple station arrangement with a standard telemetering transmitter, or a single station with a miniature transponder enable automatic storage of position data suited for digital computer use. New engineering features in the system correct automatically for variable phase errors that have plagued this type of equipment for the past decade. The simplified analogue-to-digital converter permits a low inertia system with a degree of accuracy and reliability heretofore unobtainable.

CUBIC is proud of its contributions to the armed forces which provide them with a factual basis for armament selection, and to the national air armament industry in support of their growth, and will continue to devote itself to the advancement of new techniques and their application for increasing support of the National Defense Effort.

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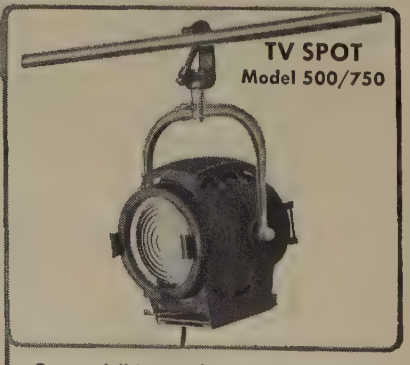
Typical Installation of B & M Spotlights in "I Love Lucy" TV Show

"Paint With Light" with

BARDWELL & McALISTER'S
Engineered Television Lights **TV SPOTS**

This complete line of television spotlights was especially designed and engineered for television stage lighting. The many years of experience in the production of studio lights for stage and motion pictures was drawn upon to produce the B & M TV Spots which make possible painting with light. Painting with light is the ability to control the light source in order to emphasize the necessary highlights and all the important shadows. Only through controlled light can the scene and subject be given the desired brilliance, beauty, and third dimensional effects to produce ideal screen pictures.



Our specialists are always ready to assist and advise your engineering staff so that your studios and stages will be fully equipped to properly paint with light. Write for complete specifications and prices of these B & M TV Spots.

BM INC

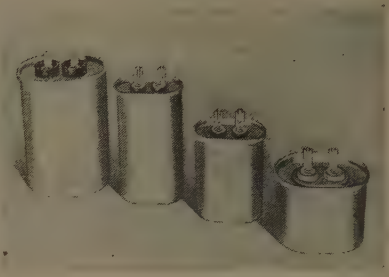
BARDWELL & McALISTER, Inc. 2950 ONTARIO STREET BURBANK, CALIFORNIA

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 140A)

Metal-Cased AC Capacitors



Cornell-Dubilier Electric Corp., South Plainfield, N. J., announces a new series of oval and flat-oval capacitors expressly designed for alternating current applications where need for space-conservation is a factor. The four new styles, which are identified as KK, KA, KD and KV (left to right in illustration), are all metal-cased paper-dielectric capacitors. Each is smaller in size and lighter than comparable rectangular shaped cases. Many newly-developed features make them suited for use in motor-starting and motor-running applications, fluorescent lighting, power-factor correction, and all general-purpose ac applications.

For complete information, write for Engineering Bulletin 156.

Jobber Catalog

Allied Radio Corp., 100 N. Western Ave., Chicago 80, Ill., distributors of radio and electronic parts and equipment, announce the release of their 1955 general catalog. The new catalog contains 308 pages listing over 25,000 items.

Special emphasis has been placed on equipment for industrial maintenance, research and production requirements. There are detailed listings of standard and special-purpose electronic tubes, test instruments, voltage stabilizers, transformers, resistors, capacitors, printed circuit components, the new transistors, rheostats, relays, switches, rectifiers, fuses, tools, wire, cable, photo-electric components, nuclear instruments, sound-powered telephones, counters, program clocks, batteries, sockets, generators, power supplies and a wide variety of other electronic equipment and components.

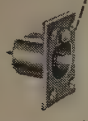
(Continued on page 144A)

In CANNON "UA" Audio PLUGS

gold-plated contacts really pan out!

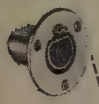
UA-3-11 Plug

UA-3-14 Wall Mounting Receptacle



UA-3-32 Wall Mounting Receptacle

UA-3-42 Microphone or Panel Mounting Receptacle



UA-3-13 Wall Mounting Receptacle

UA-3-31 Wall Mounting Receptacle

UA-3-12 Plug

Yes! You get immense satisfaction from gold in any form... and particularly from the performance of the gold-plated contacts in Cannon's modern "UA" audio Series connectors.

Contacts are heavily gold-plated over silver plate... not just "flashed" with a light coating of gold. You get maximum protection from oxidation and deterioration... maximum protection to vital audio circuits ...absolute reliability.

Cannon's thumb-pressure LATCH-LOCK prevents accidental disconnect. Available through Cannon distributors.

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Refer to Dept. 377

CANNON ELECTRIC COMPANY, 3209 Humboldt St., Los Angeles 31, Calif. Factories in Los Angeles; New Haven; Toronto, Canada; London, England. Representatives and distributors in all principal cities.

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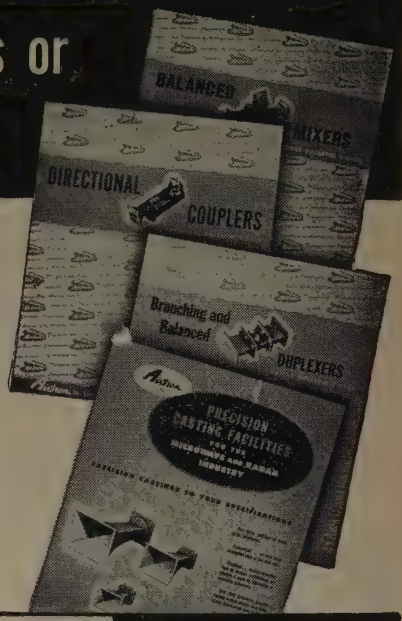
Since 1915 CANNON ELECTRIC

for SPECIFIC microwave information on "new" components or "standard" plumbing



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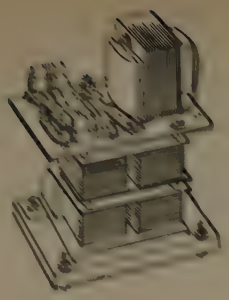
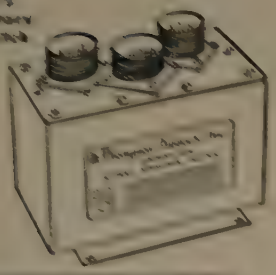


Thompson's Magnetic Amplifier

ideas...



Thompson's Airborne Rotary Selector Switch

facilities...

experience

Thompson has all three

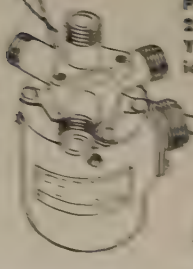
It's no accident that more and more manufacturers are turning to Thompson to solve tough electronics problems.

Thompson has ideas! Thompson engineers will not admit "it can't be done" for they are continually finding the answers to tough research, development and production problems.

Thompson has facilities! Complete development and testing laboratories, and modern production equipment are available to the skilled electronics engineers who make up the highly successful Thompson team!

Thompson has experience! For 52 years, Thompson has been blazing trails and making vital contributions to the automotive, aircraft and general industries of the nation. The highly valuable skills and experience of the entire Thompson organization are at your service for research, development and production of all things electronic.

FOR COMPLETE INFORMATION on how Thompson's Electronics Division can work for you, write to Thompson Products, Inc., Electronics Division, 2196 Clarkwood Road, Cleveland 3, Ohio. You will receive details of Thompson ideas . . . facilities . . . experience.



One of the many Thompson Coaxial Switches

Electronics Division

Thompson Products, Inc.

2196 CLARKWOOD RD., CLEVELAND 3, OHIO

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 142A)

Magnetic Plastic Welding Electrode

Incorporation of all the functions of conventional presses directly into the electrode, makes the new Ectromag Magnetic-Bar Electrode available from **Electronic Processes Corp.**, 1197 San Antonio Rd., Los Altos, Calif., useful for the welding of large areas of plastic sheeting. Electrodes of this new type have particular advantages for automatic machinery, since they facilitate simple, high-speed mechanical feeds. Plastic sheeting can be wrapped around the keeper bar for the production of unlimited lengths of plastic tubing for packaging and other requirements.

Eliminating the overhead structures of conventional presses, the Ectromag has a box-like steel magnetic structure which encloses the radio-frequency electrodes. Magnetic coils mounted below draw a free keeper bar downward when energized to apply welding pressure for the plastic sheeting against the electrode. Pressure is adjusted by adjustment of voltage through the magnet-actuating coil circuit.

Synchronized Streak Camera

With the addition of control facilities illustrated, right, the Model 398 Streak Camera manufactured by **Beckman & Whitley, Inc.**, 915 E. San Carlos Ave., San Carlos, Calif., becomes a complete system

for the synchronized recording of position-versus-time data. In use, the instrument makes a plot on 4-by-10-inch film where the vertical axis represents space and the horizontal axis represents time.

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

Transient events such as explosive, flash-tube, shock-wave and spark-discharge phenomena are suitable subjects providing they can be controlled with respect to time co-ordination. Self-luminous events, as well as schlieren and interferometric effects can be recorded.

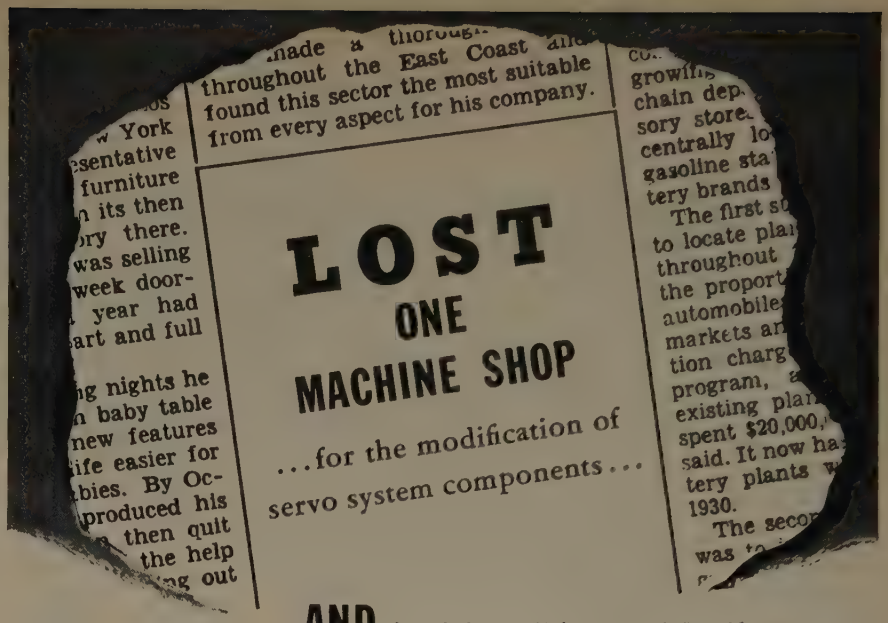
An external objective lens images the event on the camera slit. An internal lens forms an inverted, approximately 1-to-1 image of the slit on the face of a polished steel rotatable mirror. When the mirror is being driven at its maximum speed of 50,000 rpm, the image is swept across the film at a writing rate of 5.46 millimeters per microsecond, and a film-length writing time of 44 microseconds.

Decimal Counting Units

There are two types of new decimal counting units available from **Advance Electronics Co., Inc.,** 451 Highland Ave., Passaic, N. J., type 100A and Type 100B. They are direct-reading plugin electronic counters. Each unit counts the number of input pulses from 0 to 9, the 10 pulse resets the counter to 0 and at the same time produces an output pulse. The number of pulses applied to the input is indicated by the illumination of one of the ten neon bulbs installed behind a numbered plastic front panel. To increase the counting capacity, one may connect a number of these decimal counters in cascade with the output of one unit fed to the input of the next unit.

Type 100A is used for counting from 40 kc to 0 cps. Type 100B is used for counting from 100 kc to 0 cps. Both types have identical power requirements, 6.3 volts at 1.2 amperes and 300 volts at 15 ma. The input signal for both types is required to be at least 100 volts, rise time 1 μs or shorter, and duration 2 μs or longer. The resolution time is 5 μs for both types. Arrangement is provided for zero reset. The output pulses of either types are adequate to trigger another Type 100A.

(Continued on page 159A)



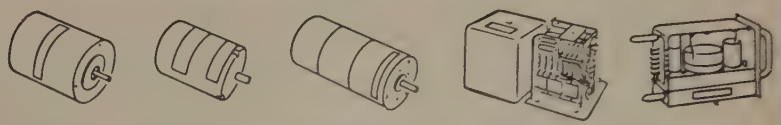
AND it might well be yours! For if you are now buying the components for servo systems from several manufacturers, you are probably wasting time, labor, machinery, and material, modifying the various units for better coordination. And you still end up with only the inferior performance that such a hodge-podge delivers.

Transicoil experience proves that you can save the time and trouble of secondary operations and end up with a better system by merely using assemblies made up of matched Transicoil components. The units comprising these assemblies are designed and constructed to work with each other for optimum efficiency, top performance, and actually cost less when assembled than the total purchase price of individual components acquired from several sources.

If you are now purchasing servo components from several manufacturers, a serious talk with Transicoil will pay you dividends in lower costs and a better system. But if you require only one component, you can be sure of optimum performance from the Transicoil units you specify.

TRANSICOIL

CORPORATION 107 GRAND STREET NEW YORK 13, N.Y.



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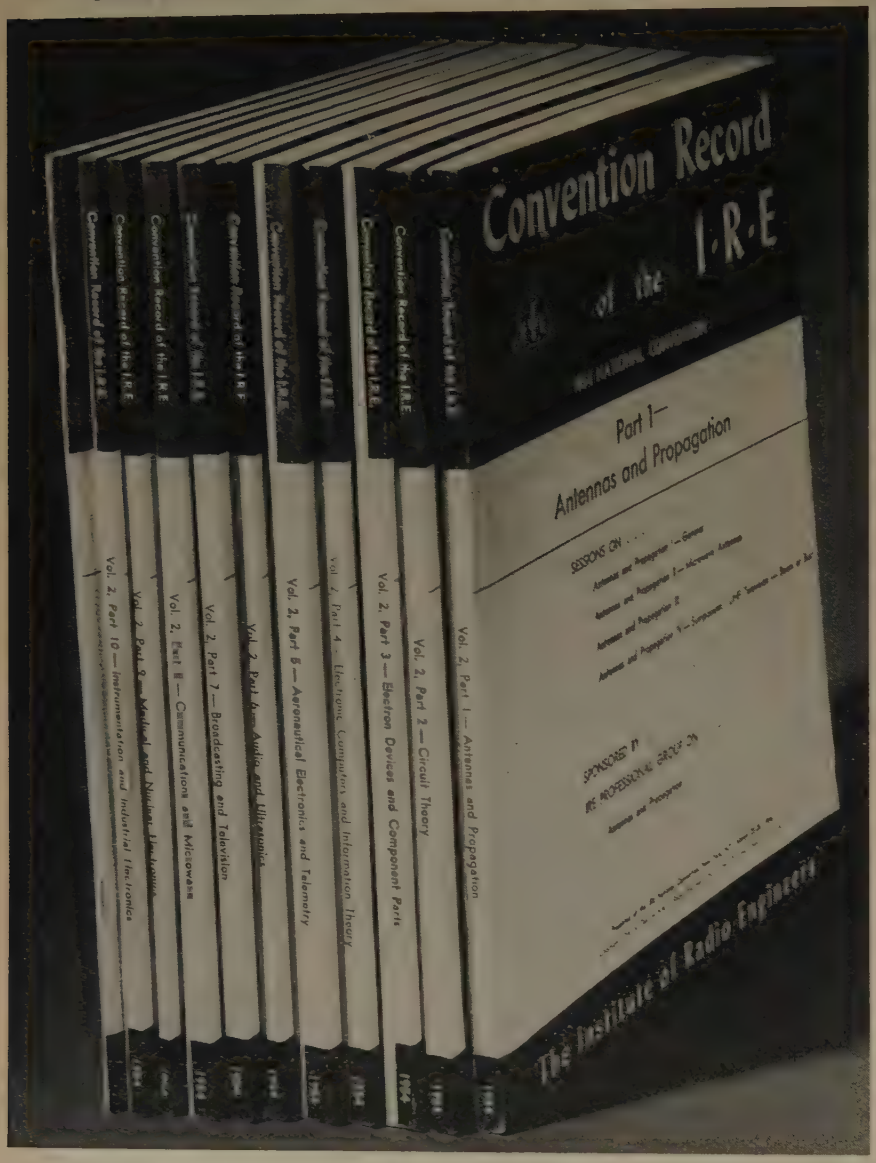
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4	Electronic Computers & Information Theory. Sessions: 2, 12, 19, 27, 34	1.50	3.60	4.50
5	Aeronautical Electronics & Telemetry. Sessions: 3, 5, 8, 10, 15, 45	1.50	3.60	4.50
6	Audio & Ultrasonics. Sessions: 11, 18, 23, 41, 48	1.50	3.60	4.50
7	Broadcasting & Television. Sessions: 13, 20, 26, 33	1.50	3.60	4.50
8	Communications & Microwave. Sessions: 1, 7, 21, 43, 50	1.50	3.60	4.50
9	Medical & Nuclear Electronics. Sessions: 17, 22, 24, 31	1.50	3.60	4.50
10	Instrumentation & Industrial Electronics. Sessions: 29, 36, 38, 44	1.25	3.00	3.75
11	Engineering Management & Quality Control. Sessions: 4, 9, 16	1.00	2.40	3.00
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St. Paul W4, Minn.

News—New Products

(Continued from page 145A)

Capacitive Level Indicator

A new Belmont level indicator is now available from **Thermo Instrument Co.**, 1310 County Rd., Belmont, Calif., capable of providing a continuous indication of level for a wide range of materials in storage or process it utilizes an element probe with no floating or moving parts nor contact-making electrodes.

Suitable for use with most non-adhesive chemicals including milk and oils, and refrigerants such as ammonia, the Freons, and all condensed gases, the instrument is operable over a temperature range of 500 to −425 F., and from high vacuum to 100,000 psi.

Various probe lengths, materials, and designs are available to cover different ranges in the rise and fall of materials, the character of substances encountered, and special problems.

Standard instruments provide for a ten-foot separation between the probe and the equipment housing. However, the interconnecting coaxial cable can, if necessary, be as long as 1,000 feet. Design is such as to permit inexpensive addition of remote indicating instruments over telemetering circuits of almost unlimited length.

Plastic Welding Press

A new Electrotherm Air Press, Model 453, has been developed by **Electronic Processes Corp.**, 1197 San Antonio Rd., Los Altos, Calif. It is capable of handling either intricate electrodes up to a platen size of 24 by 30 inches, or a bar electrode as long as 48 inches. The moving platen is supported on three guide posts 120 degrees apart. Both the top and bottom platens are fully insulated so that the live electrode can be located either above or below the work.

Plastic-welding control time is effected by an accurate inbuilt timer having a full scale of 0 to 5 seconds graduated in 1/10 seconds. Firing of the associated high-frequency generator, though timed by the timer, is controlled by a switch on the press overtravel mechanism so that rf

(Continued on page 160A)

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BS degree or equivalent experience is required. Experience in the following fields is desirable. Radar — Servomechanisms — TV — Analog Computers — Audio Systems — Instrumentation — Data Handling Systems.

Replies will be held in strict confidence. Write giving full details of education and experience to:

Mr. W. M. Hoyt, Dept. 686 (A)

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* Your future will be as big as you make it, with your advancement depending on your own merits.

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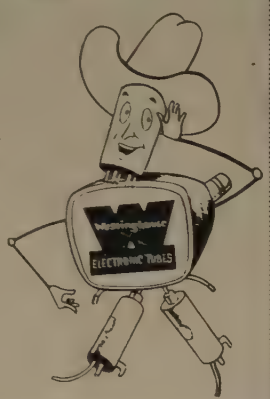
APPLICATION ENGRS: Circuit experience helpful in radio, TV, VHF, and semi-conductors.

PRODUCT ENGRS: Knowledge and/or experience in cathode design and fabrication of magnitrons and klystrons.

Recent graduates in Electrical, Chemical and Mechanical Engineering, Physics, Chemistry & Metallurgy are also urged to apply. Expenses paid if invited to Elmira for interview. Send resume:



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Systems engineers conduct studies to determine operational requirements . . . create and synthesize military equipment concepts . . . guide development of new integral elements . . . conduct evaluation programs to determine operational effectiveness.

You may qualify! Professional *ability* to create and analyze over-all complex systems and *experience* in electronic or electro-mechanical systems engineering required.

There are several opportunities now in:

AVIATION ELECTRONICS
INFORMATION HANDLING
COMPUTERS • RADAR
COMMUNICATIONS
MISSILE GUIDANCE

Send a complete resume of your education and experience to:

Mr. John R. Weld, Employment Manager
Dept. B-453K, Radio Corporation of America
Camden 2, New Jersey

RADIO CORPORATION OF AMERICA

News—New Products

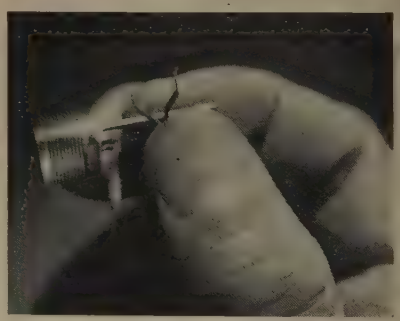
(Continued from page 159A)

power can only be applied to the electrodes when they are fully closed.

Up and down strokes of the movable platen are controlled by individual solenoid valves with separate air-metering attachments to control ram speed in both directions. A delay time is available on the up-stroke for dwell in the welding position. Positive stops control the ram position on both up and down strokes.

Weighing 72 pounds, the unit has overall dimensions of 21½ deep by 18 wide by 29 inches high, a throat depth of 15½ inches, and a daylight-opening of 7¾ inches.

Fluctuating-Pressure Transducer



Simplification and convenience in handling have been two objectives in the redesign of new bonded strain-gage transducers now being produced by **Electronic Engineering Associates, Ltd.,** 1080 Brittan Ave., San Carlos, Calif. In the new style, the reference pressure tube has been brought directly out the rear together with the electrical terminals, permitting the unit to be threaded into its flush-mounting position from either direction.

Other features are freedom from shock and acceleration effects, structural simplicity which helps to produce trouble-free service life, and ease of calibration.

Electrically, the units coordinate with the general practices of dynamic strain-gage work: usually the transducer is installed as an active arm of a 120 ohm bridge.

Standard units are ½ inch in diameter, available in two lengths: ¼ inch and ⅜ inch. Seven pressure ratings are supplied as follows: 2½, 5, 10, 25, 50, 100, and 200

(Continued on page 162A)

TUNG-SOL 50 YEARS ACHIEVEMENT

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If a year or two of practical experience has given you the youthful maturity that demands more than just a job, you may be interested in our "career opportunities" in color TV, crystal products and electronic tubes.

Submit resume or address request for personal interview to D. Bellat, Personnel Director.

TUNG-SOL ELECTRIC INC.
200 Bloomfield Avenue
Bloomfield, N.J.

ENGINEER

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•

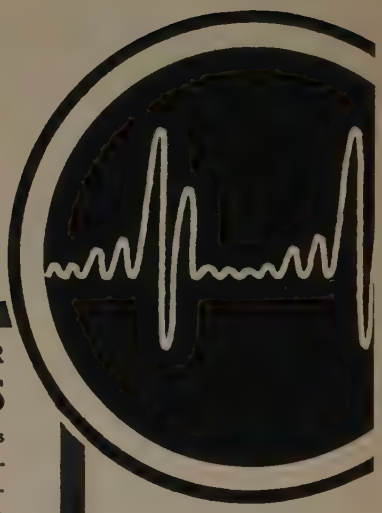
Minimum of 5 years experience specializing in design of transducers and associated equipment for high fidelity magnetic recording

•

Mail detailed resume to:
Personnel Dept., Room 1201

COLUMBIA

BROADCASTING SYSTEM, Inc.
485 Madison Ave., N.Y. 22, N.Y.



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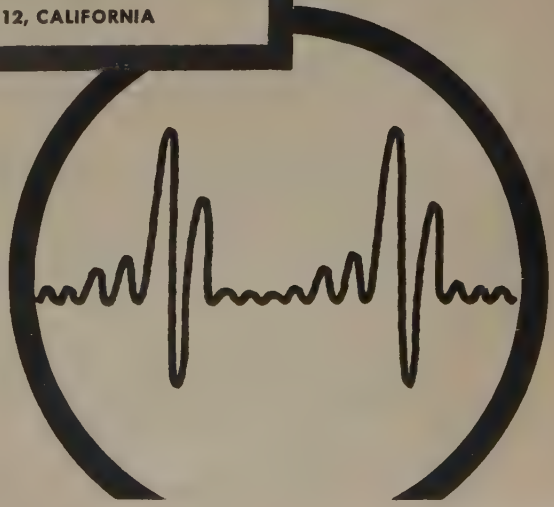

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Relocation expense allowed for those employed.

Write or Apply

DELCO RADIO DIVISION
GENERAL MOTORS CORPORATION
Kokomo, Indiana

News—New Products

(Continued from page 160A)

pounds per square inch. Linearity is 2 per cent; operating temperature range, −40 to +250° F; acceleration effect, 0.02 per cent per g; and resonant frequency of the diaphragm, 8,000 cps.

Price are $125 each for the 5, 10, 25, and 50 psi units; $135 for the other three listed ratings.

Video Probe Amplifier

The Donner Scientific Co., 2829 7 St., Berkeley 10, Calif., announces a new Video Probe Amplifier, Model 60. This amplifier is an electronic instrument designed to facilitate circuit examination with all types of oscilloscopes. Because of its low input capacity, its use with any oscilloscope permits observation of the circuit, without loading, detuning or other adverse effects on normal performance. This new video probe system introduces no attenuation, so oscilloscopes may be used at full gain for accurate observation of small signals. Model 60 Video Probe Amplifier has uniform frequency response from 5 cps to 12 mc within ±3 db, which is equal to or greater than that of most oscilloscopes.

It also serves as a general-purpose amplifier with gain of 40× and more than 12 volts peak-to-peak output. The probe of this instrument clips on and can be used as an oscilloscope accessory. Also, it has an RG/62U cable input with UHF connectors to use the instrument as an amplifier.

Price is $150.00 Fob Berkeley, Calif. Request Data Sheet 601.

FM Modulation Meter

Lampkin Laboratories, Inc., Bradenton 13, Fla., has recently introduced their 205-A FM modulation Meter. The instrument has a range of 25 to 500 mc; it is an improved version of their type 205. While the 205-A is primarily designed for 2-way radio maintenance, it has many other uses. It measures Fm deviation ±0–25 kc., with an accuracy within 10 per cent full scale on a 3-inch meter calibrated in kilocycles. It can also be used as a relative field-strength meter, and has a built-in speaker and a jack for an oscilloscope. Its size (7×12×7¼ inches deep) and weight (less than 14 pounds) make it suitable as

(Continued on page 164A)

ELECTRONIC ENGINEER

Interesting opportunity for key engineer to participate in basic development of multi-billion volt proton accelerator.

Require man with extensive experience in the 100-1000 MC UHF region. Record of achievment in advanced electronic development and demonstrated interest in research essential.

Brookhaven Laboratory is located in Suffolk County, Eastern Long Island, New York.

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Including Salary Required
To Personnel Dept.

BROOKHAVEN NATIONAL LABORATORY
Associated Universities Inc.
Upton, New York

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RADAR
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Openings exist at both senior and junior levels. BS, MS or PhD in Mathematics, Physics or EE required. Junior engineers with this educational background, without experience, will be accepted for training in the above openings.

For additional information please forward resume to:

Missile & Control Equipment Dept. C
Engineering Personnel

NORTH AMERICAN AVIATION, INC.
DOWNEY, CALIFORNIA

In the N. Y. area, please contact our representative, Mr. G. W. Benedict; 19 Rector St., Rm. 1609, New York 6, N.Y.

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Bedford, Mass.

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  Must have detail design experience in field of electronics with some general experience in radar applications and radar systems engineering. Must be capable of undertaking senior responsibility in important projects. Minimum of five years' applicable experence required.
- **SENIOR RF ENGINEER**
  Opening for a practical engineer with initiative and appropriate background. Primarily must have extensive experience in microwave techniques, with some experience and know-how in Radar Systems. Must possess keen appreciation of the RF requirements of radar systems to meet military specifications. Minimum of five years' applicable experience required.
- **SENIOR MECHANICAL ENGINEER**
  Must have experience in the design of mechanical components such as gear trains, hydraulic actuators, mechanical drives, gimbals, and related components for airborne systems. Must possess good working knowledge of machine practices, machine tools and their uses. Desirable to have knowledge of servomechanisms and electrical drives for ground and airborne systems. Minimum of five years' applicable experience required.
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  Experience in the design of one or more of the following categories required: Electro-mechanical components, small motors, hydraulic components, precision bearing design, servomechanism (hydraulic), instruments. Minimum of four years' experience required.
- **PLASTICS ENGINEER**
  Minimum of two years' experience required in physical chemistry, preferably with work on plastic materials. Duties involve the deevlopment of techniques using high temperature plastics for mounting and embedding electronic components. Must be capable of developing molds and predicting suitability of chosen plastics and finishes for electronic components.
- **ENGINEERING PHYSICIST (RADOMES)**
  Should have degree in Physics (or Electrical Engineering with proficiency in physics and mathematics). Two to five years' experience required in appropriate areas of physics and/or electronic engineering in circuit design, RF design, and related developments. Knowledge of dielectric properties of radome materials and experience or know-how in microwave techniques desirable.
- **ELECTRONICS ENGINEER**
  Must have three to five years' experience in the design and engineering of centimeter radars. Additional experience in Field or military electronics and aircraft installations desirable.
- **CIRCUIT DESIGN ENGINEERS**
  Several openings for engineers with a minimum of three years' experience in circuit design work.
- **HEAT TRANSFER ENGINEER**
  Four to five years' experience required in cooling techniques and heat transfer design of electronic equipment for use in Guided Missiles and Airborne Radar Systems.
- **DESIGN ENGINEER**
  Three to five years' experience required in Ground-to-Air Missile launching equipment and/or field experience with intimate knowledge of missile handling equipment.

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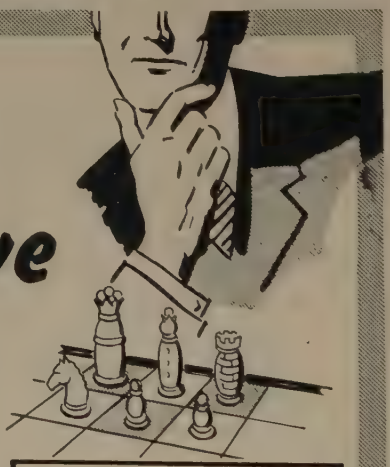
Electronic Engineers *Your Move*



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News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 162A)

either a laboratory or portable instrument. The 105-B Micrometer Frequency Meter, the companion unit to the 205-A, can now be used to measure crystal-controlled transmitters up to 500 mc.

TV Capacitors

A new high voltage ceramic capacitor with integral corona shield is being used in early model color TV receivers. This capacitor, rated at 500 $\mu\mu$F, 30,000 volts, is part of a new series of high voltage ceramics developed especially for color TV applications by the **Sprague Electric Co.**, 235 Marshall St., North Adams, Mass. Mechanical configuration of this new unit resembles a child's toy "top" in appearance.

Complete description of this new encapsulated capacitor, together with information on other units furnished by Sprague for color TV applications, are given in Engineering Bulletin 609, available upon letterhead request.

UHF Grid Dip Meter

Boonton Electronics Corp., Boonton, N. J., has announced its Model 101B UHF Grid Dip Meter. The instrument operates in the frequency range of 300 to 1000 mc. The frequency scale is approximately linear throughout the ranges which are

from 300 to 425 mc, 425 to 650 mc, and 650 to 1000 mc employing three plug-in coils mounted externally on the UHF probe. The dial is individually calibrated to a frequency accuracy of 2 per cent.

(Continued on page 167A)

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 164A)

The instrument has innumerable applications in the UHF field, such as measurement of capacity, inductance, circuit Q, choke resonance. It can be used as an auxiliary signal generator, or an absorption wave meter.

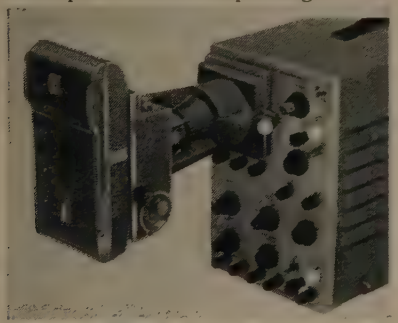
A three-position switch located on the power supply panel enables the instrument to be used in CW, modulated CW or as a detector. The catrier can be internally or externally modulated.

As a detector, the instrument can be used to determine the frequency of unknown RF signals and parasitics. It may also be used as a field strength meter.

It is in two units, a supply and a UHF probe. The probe is connected to the power supply with a 4 foot cable and plug for changing to low frequency probe which will be available at a future date.

Oscilloscope Camera

A new fully automatic oscilloscope camera, of special configuration and based on an adaptation of the Polaroid* fast-print principle, has been developed by **Aremac Associates,** 329 W. Washington St., Pasadena 3, Calif. The new Recordoscope, Model 1185, will provide an accurate and permanent photographic documentation of CRT phenomena 60 seconds after exposure of the scope image.



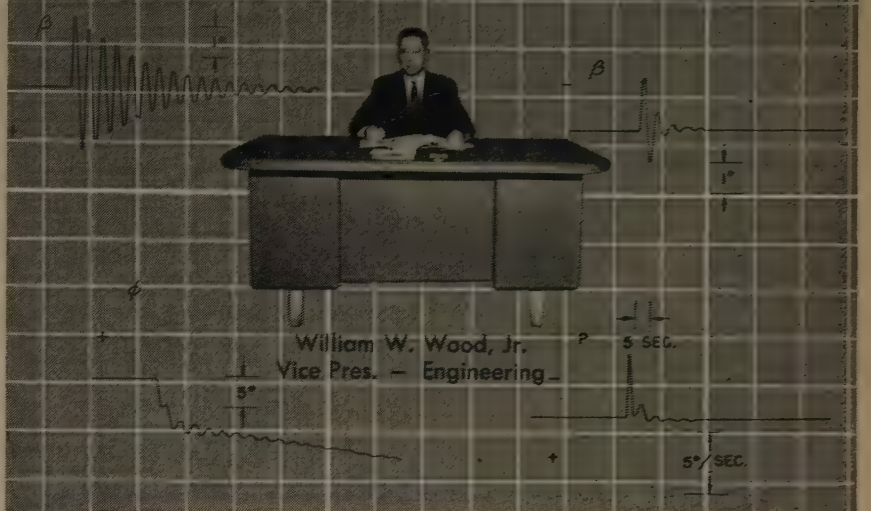
The Recordoscope takes a standard Polaroid magazine and uses a fast self-developing film. The magazine-shifting mechanism records 3 to 16 traces per 3 inch×4 inch print. Automatic cycling, exposure plus automatic movement to next trace position can be accomplished in three ways: (1) manual shutter release, (2) cable release, (3) remote operating switch. Final print provides full size image on 3 inch scopes, half-size image on 5 inch scopes, with no reversal of image.

A specially designed camera-to-scope hood is of the swing-away type and insures sturdy camera mounting. One camera can service several 3-inch and 5-inch scopes of different makes. When provided with adapter plates and hoods, camera can be moved from hood-to-hood in a matter of seconds. General Specification: Coated Oscillator f/2.3–67.5 mm lens; "time" and "bulb" shutter having speeds from 1-sec. to 1/100th second; fixed focus; Camera measures approximately $4\frac{1}{2}\times6\frac{1}{4}\times9\frac{1}{2}$ inches; Camera weight approximately 8 pounds; 3-inch hood 7 inches long, 5-inch hood $8\frac{3}{4}$ inches long.

* TM of the Polaroid Corp.

(Continued on page 169A)

SOME STRAIGHT TALK TO ELECTRONIC ENGINEERS



Here at Link Aviation, we're making substantial contributions to the field of automation. Quite possibly, there's a spot for you in our group of Senior Electronic Engineers — if you can assume full responsibility for completing electronic component or systems design problems — or can direct and coordinate the technical activities of other persons assigned to assist you.

To qualify, you must have proven ability in at least one of the following areas:

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DIGITAL COMPUTATION

Imaginative management provides an ideal working environment for our creative engineers — and hand in hand with this progressive spirit go maximum employee benefits.

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Vice President — Engineering,
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ENGINEERS

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Mechanical: Minimum of five years experience in packaging electronic equipment as well as in mechanical design of small electro-mechanical devices.

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are you EXPERIENCED in

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. . . Send us a resume of your background. We may have a higher level position open for you!

Send resume to Mr. L. H. Noggle, Dept. M-

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Division of Bendix Aviation Corp. Baltimore-4, Md.

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Living and working conditions are of the best. The modern laboratory is located in New York's Westchester County, known throughout the country for its beautiful surroundings and high standard of living . . . and only one hour from metropolitan New York city with its wealth of cultural and educational activities.

If you're interested in a permanent, satisfying career, send your resume to Mr. H. F. Ware, Personnel Director. Expenses will be paid for qualified applicants who come for interviews. We regret we can consider only U. S. citizens.

GENERAL PRECISION LABORATORY INCORPORATED

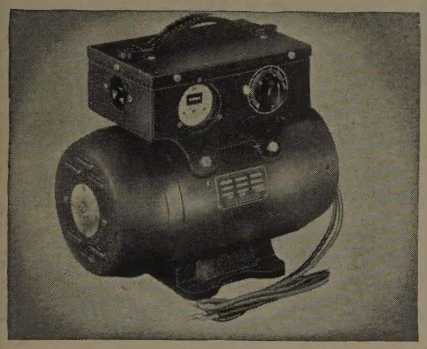
A subsidiary of
General Precision Equipment Corporation
63 Bedford Road
Pleasantville New York

News—New Products

(Continued from page 167A)

Custom Converter for Hi-Fidelity

Carter Motor Co., 2645 N. Maplewood Ave., Chicago 47, Ill., Dept. 4, has introduced a new custom converter for hi-fidelity installations. For installations that require less than 250 watts with all equipment in operation.

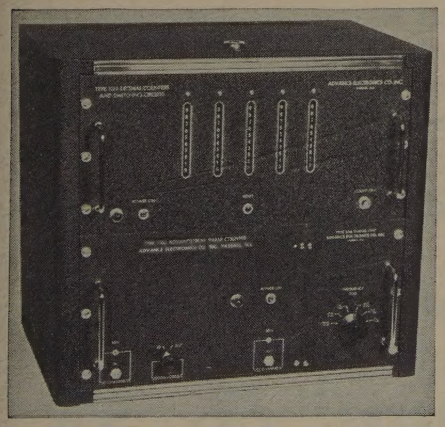


This converter is designed to deliver 125 volts ac with 120 volts dc input and a load of only 50 watts. Due to the fine regulation of the 500 watt converter, the voltage would only drop to about 105 volts with 110 volts dc input and full 250 watt load. The above ac limits coincide with equipment ratings.

A frequency control with frequency meter is included, allowing the operator to adjust exactly to 60 cps for record players and tape recorders with a filter also available to give noise-free radio reception. Units can be supplied without frequency control and/or filter, and for any input voltage from 12 to 220 dc input.

Ultra-Low Frequency Phase Counter

The outstanding features of an Ultra-Low Frequency Counter developed by **Advance Electronics Co., Inc.,** 451 Highland Ave., Passaic, N. J., are as follows: new gating circuit with interlocking arrangement offers dependable operation, plug-in units increase the usefulness to



three or four times, the lower limit of the frequency range can be extended to +0 cps with an external timing unit. This instrument consists of three plug-in units, a decade counter and switching circuit, a timing unit, and a function unit. The

(Continued on page 171A)

ELECTRONIC ENGINEERS

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WITH A LEADER IN
WESTERN ELECTRONICS

An expanding program of:

- *research*
- *development*
- *production*
- *specialized military equipment*
- *advanced commercial design*
- *real creative challenge*

Special receivers and transmitters, DF and DME, various instruments and Transistor applications—special devices. Studies in noise, radar, miniaturization and test equipment. Relocating expenses, good insurance plan, central location, steady advancement.

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Industrial Research Laboratories
Dept. B-11, Hilltop & Frederick Rds.
Baltimore 28, Maryland

Digital Computer Techniques

Applied to the design, development and application of

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Aircraft Control and Navigation Systems

Electronic Business Systems

The successful application of Hughes airborne digital computers to high speed aircraft fire control problems has opened up an entire new area for these digital computer techniques.

Similar equipment is now under development in the Advanced Electronics Laboratory to apply such digital systems to modern business information handling.

Areas include

LOGICAL DESIGN
COMPONENT DEVELOPMENT
PROGRAMMING
MAGNETIC RECORDING
CIRCUIT DESIGN
INPUT & OUTPUT DEVICES
SYSTEMS ANALYSIS

Engineers and Physicists

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Scientific and Engineering Staff

Hughes

RESEARCH AND DEVELOPMENT LABORATORIES
Culver City, Los Angeles County, Calif.

Assurance is required that relocation of applicant will not disrupt an urgent military project.

What is
Stereatronics?

General Electric Company

The "rate-grown" process of producing germanium makes it possible to obtain large quantities of uniform transistor elements with pre-specified operating characteristics. The ingot shown above takes only two hours to produce and will supply material for 2000 transistors.—Proceedings of the IRE

HARNESSING electrons in solids is opening up another stage of the bright new world that captures the imagination of men, has been given this name by Associate Editor Cornelius Ryan of COLLIER'S in his September 3rd article "Stereatronics." He specifically defines this phase of radio as "The science of controllable electronic performance of solids."

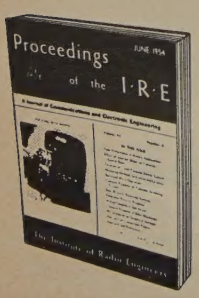
IT IS RADIO ENGINEERING!

Even before 1920, radio engineers worked skillfully with the earliest "stereatrons," glistening galena crystals that captured code and music from the skies. As years went by they developed practical uses for more "stereatrons" such as: quartz crystals engineered to vibrate at selected frequencies; phosphors that converted power to light and color on the face of video tubes; germanium particles captured in diodes and transistors that make subminiaturization and Dick Tracy's wristwatch radio practical; selenium that converts AC to DC; silicones that change light to power, which will some day heat our homes; and magnetic memory "ferrites" that almost think in computers, now opening vast vistas of industrial applications to radio.

MADE BY RADIO ENGINEERS!

As you read more about "Stereatronics" and dream of capturing new markets, remember that radio engineers have understood these semiconductors, or "stereatrons" and exchanged engineering information about them in "Proceedings of the I.R.E." years in advance of the resulting marketable products. The eleven men mentioned in Mr. Ryan's article are all radio engineers, most of them members of long standing in the IRE. Such men, because of their knowledge, "set the specs" and control the buying for the huge new manufacturing their design creates. "Stereatronics" takes form and becomes practical published information in the history making issues of

"Proceedings of the I·R·E"



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News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

Continued from page 169A)

decimal counter and switching unit has a fast gate with interlocking arrangement and a decade counter consisting of five decimal counters connected in the tandem.

The timing unit consists of 100-kc crystal oscillator and five stages of 10–1 frequency dividers. A switch is provided on the front panel to select the number of output pulses per second from this unit. The function unit consists of two new circuits known as Advancetron. One of them is used to generate a sharp pulse during the instant when the input signal, E_1, intersects with the zero-axis. The other Advancetron is used to generate a sharp pulse during the instant when the input signal E_2, intersects with the zero axis. The function unit can be pulled out and changed to another one for converting the instrument to a frequency counter, a timer interval counter, or a universal time interval and frequency counter.

COMMUNICATIONS EQUIPMENT CO.

MICROWAVE COMPONENTS

10 CM.—RG48/U Waveguide

10CM ECHO BOX: Tunable from 3200-3333 Mc. For checking out radar transmitters, for spectrum analysis, etc. Complete with pickup antenna and coupling devices. $27.50
10 CM ANTENNA ASSEMBLY: 3000-3300 Mc. Parabolic Dish, 29 inch Diam. Fed from dipole Rotation: 360 Deg. Azimuth at speeds of 20 and 10 RPM. Tilt: 20 deg. above and below horizontal. Motor-Driven by 2-28V motors. 4.5 A Total drain. Azimuth info. is fed to selsyn mechanism, and elevation data is obtained from Azimuth potentiometer. Net weight 65 lbs. $78.50
POWER SPLITTER for use with type 726 or any 10 CM Shepherd Klystron. Energy is fed from Klystron antenna through dual pick-up system to 2 type "N" connectors .....$22.50 EACH
LHTR, LIGHTHOUSE ASSEMBLY. Parts of RT39 APG 5 & APG 15, Receiver and Trans. Cavities w/assoc. Tr. Cavity and Type N CPLG, To Recvr. Uses 2C40, 2C43, 1B27, Tunable APX 2400-2700 MCS. Silver Plated $22.50
BEACON LIGHTHOUSE cavity p/o UPN-2 Beacon 10 cm. Mfg. Bernard Rice, each $27.50
MAGNETRON TO WAVEGUIDE Coupler with 721-A Duplexer Cavity, gold plated $45.00
721A TR BOX complete with tube and tuning plungers $12.50
McNALLY KLYSTRON CAVITIES for 707B or 2K28, 2700-2900 Mc. $4.00
WAVEGUIDE TO ⅞" RIGID COAX "DOORKNOR" ADAPTER CHOKE FLANGE SILVER PLATED BROAD BAND $32.50
AS14A AP-10 CM Pick up Dipole with "N" Cables $4.50
HOLMDELL-TO-TYPE "N" Male Adapters, W. E. #D167284 .....$2.75
I.F. AMP. STRIP: 30 MC, 30 d.b. gain, 4 MC Band-width, uses 6AC7's—with video detector, A.F.C. less tubes $24.00
BEACON ANTENNA, AS31/APN-7 in Lucite Ball. Type "N" feed $22.50
ANTENNA, AT49A/APR: Broadband Conical, 300-3300 MC Type "N" Feed $12.50
"E" PLANE BENDS, 90 deg. less flanges ...$7.50

3 CM.—RG 52/U Waveguide

3 CM ANTENNA ASSEMBLY: Uses 17" parabaloid dish, operating from 24 vdc motor. Beam pattern: 5 deg. in both Azimuth and elevation. Sector Scan: over 160 deg. at 35 scans per minute Elevation Scan: over 2 deg. Tilt: Over 24 deg. $85.00
Cross-Guide Directional Coupler, UG-40 output flange. Main Guide is 6" Long, with 90 Deg. "E" Plane bend at one end, and is fitted with Std. UG 39/UG 40 flanges. Coupling figure: 20 db Nominal $22.50
FLEX. WAVEGUIDE SECTION, 1 ft. long. With UG-40/UG-39 flanges. Attenuation is less than 0.1 db. at 9375 mc, and VSWR is less than 1.02. Rubber covered $7.50

JAN WAVEGUIDE FLANGES

UG 39/U	$1.10	UG 51/U	$1.65
UG 40/U	$1.25	UG 52/U	$3.40
UG 40A/U	$1.85	UG 52A/U	$3.40

VSWR Measuring Section. Consisting of 6" straight section, with 2 pick-up, Type "N" Output Jacks. Mounted ⅛ Wave apart $7.50
1" x ½" waveguide in 5' lengths, UG 39 flanges to UG40 cover. Silver plated .....per length $5.00
Rotating-joints supplied either with or without deck mounting. With UG40 flanges each, $17.50
Bulkhead Feed-thru Assembly $15.00
Pressure Gauge Section 15 lb. gauge and press nipple $10.00
Pressure Gauge, 15 lbs. $2.50
Directional Coupler, UG-40/U Take off 20db $17.50
TR-ATR Duplexer section for above $8.50
Rotary joint choke to choke with deck mounting $17.50
90 degree elbows. "E" plane 2½" radius ..$12.50
Microwave Receiver, 3 CM. Sensitivity: 10-13μ Watts. Complete with L.O. and AFC Mixer and Waveguide Input Circuits, 6 I.F. Stages give approximately 120 DV gain at a bandwidth of 1.7 MC. Video Bandwidth; 2 MC. Uses latest type AFC circuit. Complete with all tubes, including 723A/B Local Oscillator $175.00
ADAPTER, waveguide to type "N", UG 81/U, p/o TS 12, TS-13, Etc. $14.50
ADAPTER, UG-163/U round cover to special btl. Flange for TS-45, etc. .....$2.50 ea.

1¼" x ⅝" WAVEGUIDE

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BI-Directional Coupler, Type "N" Takeoff 25 db. coupling $22.50
BI-Directional Coupler. UG-52. Takeoff 25 db. coupling $24.95
Waveguide-to-Type "N" Adapter. Broadband $17.50

MAGNETRONS

Type	Peak Range (MC)	Peak Power Out (KW)	Duty Ratio	Price
2J21A	3345-9405	50		$ 8.75
2J22	3267-3333	265		7.50
2J26	2992-3019	275	.002	7.49
2J27	2965-2992	275	.002	19.95
2J29	2914-2939	275	.002	44.95
2J31	2820-2860	285	.002	24.50
2J32	2780-2820	285	.002	28.50
2J38*	3249-3263	5		16.50
2J39*	3267-3333	8.7		24.50
2J48	9310-9320	50	.001	24.50
2J49	9000-9160	50	.001	59.50
2J56*	9215-9275	50	.001	132.50
2J61†	3000-3100	35	.002	34.50
2J62†	2914-3010	35	.002	34.50
3J31	24-27KMC	50	.001	85.00
4J34	2740-2780	900		125.00
4J38	3550-3600	750	.001	169.45
4J42†	670-730	30	.003	169.50
5J23	1044-1056	475	.001	49.00
700B	690-700	40	.002	22.50
700D	710-720	40	.002	39.75
706EY	3038-3069	200	.001	32.50
706CY	2976-3007	200	.001	32.50
725-A	9345-9405	50	.001	Write
QK60†*	2845-3005	.100 CW		85.00
QK61†*	2975-3170	.100 CW		85.00
QK62†*	3135-3330	.100 CW		85.00
QK259†	2700-2900	800	.001	249.00

*—Packaged with magnet.
†—Tunable over indicated range.

KLYSTRONS

723A	$12.50	2K25/723A/B	$27.50
723A/B	19.50	417-A (West'Hse)	17.50

VARISTORS

D-167208	$1.35	D-171812	$1.63
D-171858	$1.42	D-172155	$1.50
D-168687	$1.35	D-167176	$1.25

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D-164699 Bead Type DCR: 1525-2550 Ohms @ 75 Deg. F. Coefficient: 2% Per. Deg. Fahr. Max. Current 25 MA AC/DC $2.50
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D-167613 Disk Type DCR: 355 Ohms @ 75 Deg. F.P.M. 2.5%, 1 Watt $1.35
D-166228 Disk Type 7120 Ohms @ 60°F, 4220 Ohms @ 80°F. 2590 Ohms @ 100°F., 1640 Ohms @ 120°F $1.35

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352-7102	6.3V/2.5A ...	1.45
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702724	9800/8600 @ 32MA ...	8.95
K59584	5000V/290MA, 5V/10A ...	22.50
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352-7070	2x2.5V/2.5A (2KV TEST) 6.3V/ 2.25A, 1200/100/750V. @ .005A...	7.45
352-7196	1140/1.25MA, 2.5V/1.75A. 2.5V/ 1.75A—5KV Test ...	3.95
352-7176	320VCT/50MA, 4.5V/3A, 6.3VCT/ 20A, 2x6.3VCT/6A ...	4.75
RA6400-1	2.5/1.75A, 6.3V/2A—5KV Test ...	2.39
901692	13V 9A ...	2.49
901699-501	2.77V @ 4.25A ...	3.45
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Ux8855C	900VCT/.067A, 5V/3A ...	3.79
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KS 9445	592VCT/118MA, 6.3V/8.1A, 5V/2A	5.39
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352-7069	2-2.5V Wdgs at 2.5A, Each Lo-Cap., 22Kv Test ...	5.95
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RAYTHEON WX 4298E: Primary 4KV., 1.0 USEC. SEC: 16KV-16 AMP DUTY RATIO: .001 400 CYCLE FIL. TRANS. "BUILT-IN" $42.50
WECO: KS 9948: Primary 700 ohms; Sec: 50 ohms. Plate Voltage: 18 KV. Part of APQ-13 .....$12.50

GE #K-2449A
Primary: 9.33 KV, 50 ohms Imp.
Secondary: 28 KV, 450 ohms.
Pulse length: 1.0/5 usec @ 635/120 PPS, Pk Power Out: 1,740 KW
Bifilar: 1.5 amps (as shown) ...$62.50

GE #K-2748-A. 0.5 usec @ 2000 Pps. Pk. Pwr. out is 32 KW impedance 40:100 ohm output. Pri. volts 2.3 KV Pk. Sec. volts 11.5 KV Pk. Bifilar rated at 1.3 Amp. Fitted with magnetron well $39.50
K-2745 Primary: 3.1/2.8 KV. 50 ohms Z. Secondary: 14/12.6 KV 1025 ohms Z. Pulse Length: 0.25/1.0 usec @ 600/600 PPS, Pk. Power 200/150 KW. Bifilar: 1.3 Amp. Has "built-in" magnetron well $42.50
K-2461-A. Primary: 3.1/2.6 KV—50 ohms (line). Secondary 14/11.5 KV—1000 ohms Z. Pulse Length: 1 usec @ 600 PPS. Pk. Power Out: 200/130 KW. Bifilar: 1.3 Amp. Fitted with magnetron well .$39.75
K35J45—Pulse Inversion: PRI: 5 KV PK. Pulse Negative. Sec: Pos. Pulse, 4 KV; 1 usec. and .001 DUTY RATIO $6.50
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Ray UX 7896—Pulse Output Pri, 5v sec. 41v $7.50

TEST EQUIPMENT

TSX-4SE ANALYZER

3 cm. Spectrum analyzer, similar to TS 148, but with greater sensitivity. Freq. range: 8500-9600 MC. Features: Absorption wavemeter, waveguide-Beyond-cutoff attenuator, 20 mc I.F. Strip, 120 db gain. Operates from 115V/220 V. 50-1200 CPS .
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